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STUDY OF SOLID-STATE INTEGRATED MICROWAVE CIRCUITS

Scientific Report No. 2

U8-811500-8

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TEXAS INSTRUMENTS
INCORPORATED

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ABSTRACT

A study of basic frequency-modulation telemetry transmitter configurations suitable for use in the 1- to 2-GHz frequency range is presented. The purpose of this study is to establish the framework within which a detailed analysis of the system requirements in terms of components, techniques, and devices can be made with the objective of demonstrating the system performance when implemented in integrated circuitry. The study includes the objective specifications, bandwidth determination, AFC control system parameters, and a discussion of eleven basic configurations.

PREFACE

A study of Solid-State Integrated Microwave Circuits, under the sponsorship of the Electronics Research Center of the National Aeronautics and Space Administration, is being performed by Texas Instruments Incorporated under Contract NAS 12-75. The objective of this contract is to perform the analytical study of solid-state integrated microwave circuits, techniques, and components necessary to accurately define the problem areas associated with integrated circuits when various combinations of active and passive circuit elements are required to perform a complete circuit function at microwave frequencies.

In pursuance of this objective, this report presents the results of work performed under the second of four items of the work statement. The period covered is 15 December 1965 through 14 March 1966. This second task is concerned with the selection of basic FM telemetry transmitter configurations capable of demonstrating the problems associated with microwave integrated circuits at particular frequencies in the region between 1 GHz and 2 GHz.

Separate sections of the report present discussions of the objective specifications, the bandwidth requirements, the automatic-frequency-control system parameters, the indirect and direct FM techniques, and an analysis of eleven basic direct FM transmitter configurations.

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

LIST OF ILLUSTRATIONS (Continued)

LIST OF TABLES

SECTION I

INTRODUCTION

In recent years, through application of the integrated circuit, great strides have been made in digital systems designed for the space environment. Improvements have been made in performance, reliability, reduction of size and weight, and reduced primary power requirement. These improvements now appear to be feasible in the microwave frequency range through the use of integrated circuitry techniques specifically designed for this frequency range.

As a first step in establishing this feasibility, a study was performed of solid-state microwave devices, techniques, and components suitable for use in the 1- to 6-GHz frequency range with application to integrated circuitry. This three-month study was completed 14 December 1965. The work reported here covers a second three-month study task concerned with the selection of basic FM telemetry transmitter configurations suitable for use in the 1- to 2-GHz range and for implementation in integrated circuits. It is essential that such a study be made, since we are dealing with a new set of building blocks and a new set of constraints. Conventional approaches may no longer be applicable and approaches that were rejected in the past due to component problems may prove superior when consideration is given to the new components, techniques, and devices.

The performance requirements for this FM telemetry transmitter are those normally associated with earth-space telemetry applications:

Frequency	1.0 to 2.0 GHz
Frequency stability	±0.005%
Modulation	
Type	FM and/or FSK
Linearity	1% for FM
Baseband	2 KHz to 1 MHz
Sensitivity	0.2 volts rms/100 KHz
Deviation	±1.5 MHz
Input impedance	600 ohms
Output impedance	73 ohms
Power output	1.0 to 5.0 watts
DC input	−24.5 volts

Efficiency (dc to RF)	10% to 20%
Size	Approx 6" × 2" × 6-1/2"
Weight	Approx 3 to 4 lb

In this study, following an analysis of the fundamental considerations involved, a relatively large set of basic block diagrams of FM telemetry transmitter configurations have been investigated. Those systems that were clearly inadequate with regard to meeting the performance specifications listed above were eliminated. Also eliminated were those systems fundamentally capable of meeting the specifications but having objectionable performance characteristics. There was no attempt to prejudge the circuitry investigations to be performed in the next phase of this study program, and for that reason four basic approaches to the design of the transmitter have been selected for further analysis in the next phase. These are reviewed in the summary of Section II.

SECTION II

SUMMARY OF RESULTS

Basic FM transmitter configurations may be classified according to the method of generating the frequency modulation—direct or indirect—and, if direct, according to the method of automatic frequency control (AFC). The indirect method employs phase modulation and requires that the modulating signal be integrated so that frequency modulation may be produced by the phase modulator. Since only a small amount of phase deviation can be obtained in the modulator, a large amount of frequency multiplication is needed to produce the required frequency deviation. In the specific case under investigation, multiplication ratios on the order of 2500 are needed. The problems of obtaining this amount of multiplication and the filtering requirements associated with the spurious signal suppression lead immediately to the rejection of the indirect method of frequency modulation in favor of the direct method.

Several of the direct FM techniques are not capable of meeting the performance requirements. The voltage-controlled crystal oscillator (VCXO), for example, does not allow deviation rates much greater than 100 KHz, which is far below the 1 MHz specified. Thus, although this system is conceptually simple and would be an ideal solution to the problem, it is fundamentally incapable of meeting the requirements. Another example is the simple discriminator AFC system. Because of the poor center-frequency stability of the discriminator (about ±0.25 percent long term), the center frequency of the discriminator must be low in order for the output frequency to be maintained within ±0.005 percent. On the other hand, because of the poor stability of the uncompensated voltage-controlled oscillator (VCO), the bandwidth of the discriminator must be wide. The combined requirement dictates the need for a discriminator having a bandwidth equal to approximately twice the center frequency; this is, of course, impracticable. Still another example is the heterodyne technique wherein a stable crystal-controlled-oscillator frequency is combined with the unstable VCO frequency. The ratio of the stable frequency to the unstable frequency is made large to minimize the effect of the instability of the VCO on the output frequency. In Section IV.D, it is shown that the combined requirements of modulating spectrum bandwidth and the ratio of VCO center frequency to deviation frequency—even for poor modulation linearity—cause the output center frequency stability to be greater than ±0.005 percent.

Two other systems have been eliminated; although they are fundamentally capable of meeting the performance specifications, their AFC characteristics exhibit regions of positive feedback. These two are called the dual-oscillator single-mixer AFC system and the dual-oscillator gated AFC type II system; they are discussed in Sections IV.H and IV.J respectively. A pulse-discriminator system discussed in Section IV.F has been eliminated for further investigation because its operation requires a variable rate pluse train, the rate of which falls within the modulation frequency range. It would probably be very difficult to eliminate noise in the modulation spectrum because of this pulse train. Some other systems considered employ some form of pulsed operation, but the pulse frequencies used are above the modulation frequency range, thus making it much easier to prevent them from reaching the modulator input.

Four of the systems studied are recommended for further study in the next phase of the contract. All four are capable of meeting the fundamental specifications, pending specific circuitry investigations. Two of these require RF switching; for example, either two reference frequencies are alternately gated into a mixer or a reference frequency and the VCO output frequency are alternately gated into a discriminator at a low frequency rate, considerably below the lowest modulation frequency. The other two systems require no switching, and two of these AFC systems are basically linear signal systems, whereas the other two are pulse systems. Thus, the four systems encompass a relatively wide range of techniques and components; one or more should prove very useful. The four systems are discussed briefly in the remainder of this section.

The gated discriminator AFC system (Section IV.B and Figure 12) functions by alternately sampling the output of the VCO and a crystal-controlled reference frequency. The samples are fed to the input of a limiter-discriminator circuit. The output of the discriminator is ac coupled, which eliminates any dc offset caused by the center frequency of the discriminator. In this way the performance of the AFC loop is made independent of the discriminator center-frequency stability. The peak-to-peak value of the derived ac output is proportional to the difference between the average frequency of the VCO and the crystal-controlled reference frequency. This ac signal is next synchronously detected (using the same gating signal applied to the discriminator input switch) and filtered to provide a control signal for the VCO. This system requires the switching of S-band signals, but this is quite feasible in microwave integrated circuitry due to the recent development of PIN surface oriented diodes suitable for use in microstrip integrated circuits. Furthermore, an approach to an S-band microstrip discriminator is available and will be investigated shortly. The gated system is, however, not without its drawbacks, the principal one being the possibility of generating beats between the lower modulation frequencies and harmonics of the switching waveform. These unwanted frequencies appear in the modulated output. Better control of the high-frequency response (in the region below the normal baseband) of the AFC loop would allow these beats to be held below 60 dB in the modulated output. This control can be provided by using a more complex transfer function for the low-pass filter than that assumed in the analysis, a simple 20 dB per decade rolloff. Recent advances in the design and fabrication of thermally coupled devices make this type of transfer function possible for very low frequencies in a very small space.

The quadrature AFC system (Section IV.E and Figure 22) requires no RF switching. A 90-degree phase shift is needed, however, and this is readily provided by S-band microstrip branch line couplers over a frequency range broad enough to cover a complete telemetry band. This system uses in-phase and quadrature reference signals mixed with the VCO output to produce quadrature difference frequency signals. One of these signals is then differentiated to produce an output whose amplitude is proportional to the difference between the VCO frequency and the reference frequency and whose phase has been shifted back by 90 degrees. This signal, when synchronously detected by the signal from the in-phase channel, is provided with the sign of the frequency difference. On low-pass filtering, a control signal for the VCO is obtained. Though the differentiator would normally be the problem in this system, integrated-circuit operational amplifiers can be used in this application. Since their noise level when operated as differentiators with a bandwidth of 100 MHz is 90 dB below the maximum output, they appear to offer an obvious solution to the problem. These circuit problems will be investigated in detail in the next phase of this program.

The dual oscillator AFC system (Section IV.G and Figure 24) is conceptually simple. Two mixers and two crystal-controlled oscillators are used. The two reference frequencies are equally displaced above and below the desired output frequency such that the difference frequencies out of the mixers are the same when the VCO is exactly on the assigned frequency. For other VCO frequencies, the difference between the two difference frequencies is equal to the displacement of the VCO from the assigned frequency. The AFC signal is developed by an analog type of pulse counting discriminator. This system requires two crystals, which is a disadvantage since these are very large components when compared to the integrated circuits. Also, two frequency multipliers and two mixers are needed along with balanced limiter, differentiator, detector chains in the two sides of the system. In spite of the additional circuitry, compared to other systems, this one may prove to be easy to implement and for that reason was not eliminated at this stage.

The last system to be further evaluated in the next phase of the program is the dual-oscillator gated AFC, type I system (Section IV.I and Figure 28). It is likely that one of the multipliers shown in the block diagram (Figure 28) can be eliminated and the switch be used to alternately gate the two oscillator outputs into a single multiplier. This gated system eliminates the requirement for balanced limiter, differentiator, detector chains, since one chain is used for both of the difference frequencies. RF switching will be considered here, but it is not actually required, particularly if switching is accomplished at the oscillator outputs; in that case gated amplifiers may be used. This system is interesting, but the comments on spurious response discussed in connection with the gated discriminator AFC system also apply here; the requirement for two crystals, both of which must be changed with changes in assigned frequency, is a slight disadvantage.

All four of these systems offer a potential solution to the AFC problem. Circuitry investigations to be conducted shortly will determine which system is most adaptable to implementation in microwave integrated circuitry.

SECTION III

FUNDAMENTAL CONSIDERATIONS

A. BANDWIDTH REQUIREMENTS

One of the difficult parameters to specify for an FM system is the bandwidth required for circuitry that must pass the modulated carrier. While the importance of this parameter is not as great in transmitter design as it is in receiver design—particularly for the integrated circuitry engineer who tries to use broadband circuits as much as possible to facilitate design and manufacture—it is still significant. The AFC circuitry, for example, is an area where this information is needed in an FM transmitter and, on occasion, overlooked. For instance, if a discriminator is used in the AFC, reasonable linearity must be maintained over the bandwidth of the modulated carrier; otherwise the average output of the discriminator will not be linearly related to the average frequency of the FM wave. In all circuitry through which the modulated wave passes before transmission, it is essential that a flat amplitude characteristic and a linear phase characteristic be maintained. The objective is to have the spectrum of the transmitted wave determined by the baseband signal and the frequency deviation and in no way be influenced by the circuitry through which it passes.

Both FM and/or FSK modulation are required. For FM the instantaneous frequency of the modulated wave differs from the carrier frequency by an amount proportional to the instantaneous value of the modulating wave. The degree of nonlinearity allowed for this process is 1 percent. For FSK the modulating wave shifts the output frequency between predetermined values corresponding to the frequencies of correlated sources. Thus, FSK can be accomplished by switching between the outputs of two multipliers driven from the same oscillator, providing the frequencies used for the one and zero are integral multiples of the data rate and are phase-locked to the data rate. The disadvantage of this approach to FSK is that premodulation filtering of the data signal to minimize spectrum spreading cannot be employed, and also, an FSK modulator of this type cannot be used as a linear FM modulator. On the other hand, if FSK is accomplished with a linear FM system (PCM/FM), premodulation filtering can be employed. A linear FM modulator will allow modulation by a baseband consisting of a group of modulated subcarriers, or a broadband analog video signal, or a data signal. For the present application, a linear FM modulator capable of providing both FM and PCM/FM is the obvious solution. In the following material we consider the bandwidth requirements for both of these cases.

1. FM

For linear frequency modulation the parameters of interest as given in the specifications are the peak frequency deviation, 1.5 MHz, and the bounds on the modulating baseband, 2.0 KHz and 1.0 MHz. A simple and conservative approach often used to determine the bandwidth is a Fourier expansion consisting of writing the equation for the modulated waveform and assuming sinusoidal modulation by a single frequency component. This approach yields the familiar group of sidebands symmetrical about the carrier frequency and occurring at harmonics of the modulating frequency. The amplitudes of these sidebands are determined by Bessel functions of order corresponding to the order of the sidebands with an argument equal to the modulation index, defined as the ratio of the peak frequency deviation ΔF to the maximum modulating frequency f_h. For this specific case the modulation index is 1.5. Table I lists the amplitudes for all sidebands greater than 1 percent for a modulation index of 1.5.

Table I. Amplitudes of Significant Sidebands in a Frequency-modulated Wave for a Modulation Index of 1.5

$J_o(1.5)$ Carrier	$J_1(1.5)$ 1st Order Sidebands	$J_2(1.5)$ 2nd Order Sidebands	$J_3(1.5)$ 3rd Order Sidebands	$J_4(1.5)$ 4th Order Sidebands
f_o	$f_o \pm f_h$	$f_o \pm 2f_h$	$f_o \pm 3f_h$	$f_o \pm 4f_h$
0.5118	0.5579	0.2321	0.0610	0.0118

From a rule-of-thumb criteria normally applied, the bandwidth should be wide enough to include all components having amplitudes greater than 1 percent. On this basis the bandwidth should be $\pm 4f_h$; with an f_h of 1 MHz the bandwidth would be 8 MHz. The reason this creteria yield a large bandwidth is simply that the highest frequency component f_h is only one component in the modulating wave and as such is not capable of producing the peak frequency deviation ΔF.

To better understand the power spectrum of the modulated wave, we first recognize that the baseband typically consists of a number of modulated subcarriers or a video signal and not a single sinusoidal component at the top end of the baseband. Such a modulating signal is made up of a large number of components of varying amplitudes, frequencies, and phases. Accordingly, the modulating signal will have, on the average, essentially the same characteristics as band-limited white gaussian noise. For this reason, we can use the results obtained by Medhurst for the spectral distribution of a carrier frequency modulated by a baseband of noise. [1]

The curves of Figure 1 are taken from Medhurst's Figure 1. The only conversion necessary is in the frequency deviation: in the case of a noise baseband an rms frequency deviation is specified, whereas in our case, peak frequency deviation is specified. If we consider the peak deviation to be two times the rms deviation (the two-sigma value of the normal distribution), the peak value will be exceeded 4.54 percent of the time; if we consider it to be three times the rms deviation (the three-sigma value), the peak value will be exceeded only 0.26 percent of the time.

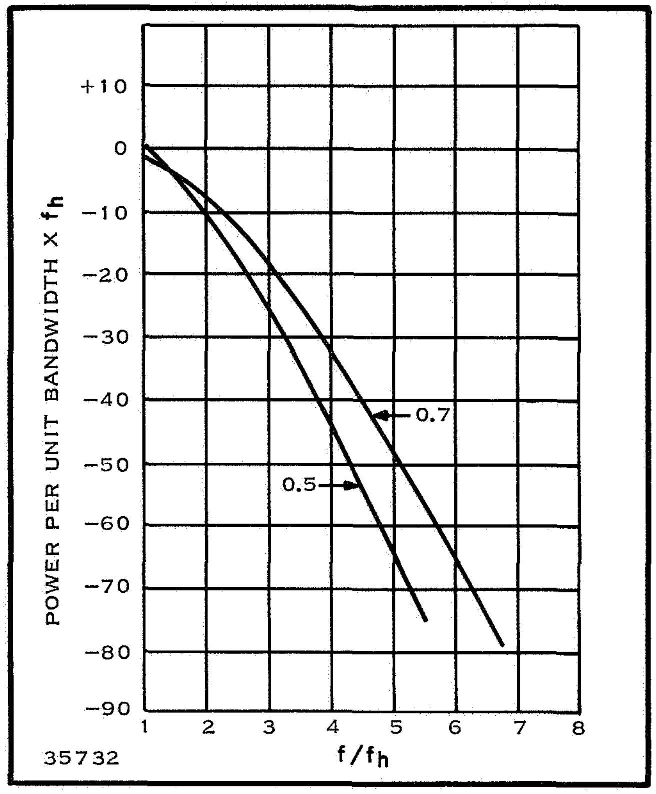

Figure 1. Spectral Distribution of a Carrier Frequency Modulated by Noise for a Ratio of RMS Frequency Deviation to Maximum Modulating Frequency, f_h, of 0. 5 and 0. 7

For a peak frequency deviation of 1. 5 MHz, the two-sigma case corresponds to an rms deviation of 0. 75 MHz and the three-sigma case corresponds to an rms deviation of 0. 50 MHz. With the maximum modulating frequency specified as 1. 0 MHz, the ratio of the rms deviation to the maximum modulating frequency for the two cases is 0. 75 and 0. 50.

The power spectral densities for these two cases are shown in Figure 1. (There is a negligible difference in the densities for the deviation ratios of 0. 7, one of Medhurst's cases, and 0. 75, one of our cases.) With the curves of Figure 1, a graphical integration of the tails of the spectrum was performed for determining the bandwidth containing all but 0. 1 percent of the power in the modulated waveform. For the two cases investigated, the required RF bandwidth (Table II) is $\pm 2.0 f_h$ and $\pm 2.6 f_h$.

For the purpose of establishing useful design criteria, we will require that all circuitry through which the

Table II. Bandwidth Containing 99. 9 Percent of the Power of a Carrier Frequency Modulated With White Noise Having Maximum Modulating Frequency* f_h

$\frac{\Delta F_{rms}}{f_h}$	$\frac{\Delta F_{peak}}{f_h}$	Bandwidth
0. 5	$\left(\begin{array}{c}1.5\\ \Delta F_{peak}/\Delta F_{rms} = 3\end{array}\right)$	$\pm 2.0 f_h$ = 4.0 MHz
0. 75	$\left(\begin{array}{c}1.5\\ \Delta F_{peak}/\Delta F_{rms} = 2\end{array}\right)$	$\pm 2.6 f_h$ = 5.2 MHz

*f_h = 1. 0 MHz

modulated wave passes before transmission meet the conservative requirement stated in the discussion of Table I; for example, the bandwidth should be $\pm 4f_h$, or 8.0 MHz. This will ensure negligible distortion of the transmitted wave. However, for circuitry used only in the AFC loop, where the only requirement is for some device (for example, a discriminator) to have an average output linearly related to the average RF frequency, we may choose the more conservative of the bandwidths derived from considering the carrier to be modulated with a noise baseband. Thus, the bandwidth of these circuits must be broad enough to cover the $\pm 2.6f_h$ or 5.2 MHz plus additional bandwidth to allow for oscillator tolerance and drifts.

2. PCM/FM

The PCM/FM power density spectrum for a large ratio of carrier frequency to bit rate (this holds in our specific case) has been derived in closed mathematical form by assuming a random sequence of binary FM signals in the form of two equally probable frequencies with continuous phase transition.[2] From this work, computations of the bandwidth required to contain specified percentages of the total spectral power have been made.[3] Figure 2 is a typical example of the results of these computations. The effect of premodulation filtering has been considered using an approximation.[4]

For optimum detection of PCM/FM, deviation ratios on the order of 0.7 to 0.8 are used, where deviation ratio is defined as the peak-to-peak deviation divided by the bit rate. For this specific case, and making full use of the 1.5 MHz peak frequency deviation, data rates on the order of 3.75×10^6 to 4.28×10^6 bits per second would produce the maximum bandwidth requirement. For lower data rates, the peak deviation normally would be reduced. For digital transmission, it is sufficient to specify that the RF bandwidth for the transmitter shall include 99 percent of the power in the modulated wave. Curves based on work previously referenced were used to obtain the results of Table III.

Table III. Bandwidth Containing Specified Percentages of the Power in a Carrier Frequency Modulated With a Random Binary Data Sequence With and Without Premodulation Filtering

Percent of Power	Without Premodulation Filter	With Premodulation Filter*
	Deviation Ratio = 0.7 $(2\Delta F_{peak}/f_b)$ $f_b = 4.28 \times 10^6$ bits per second	
90	$0.97\ f_b = 4.16$ MHz	$0.96\ f_b = 4.12$ MHz
99	$1.79\ f_b = 7.67$ MHz	$1.60\ f_b = 6.86$ MHz
	Deviation Ratio = 0.8 $f_b = 3.75 \times 10^6$ bits per second	
90	$1.01\ f_b = 3.79$ MHz	$1.01\ f_b = 3.79$ MHz
99	$1.94\ f_b = 7.27$ MHz	$1.78\ f_b = 6.67$ MHz

*Gaussian low-pass filter with 3-dB cutoff at the bit rate

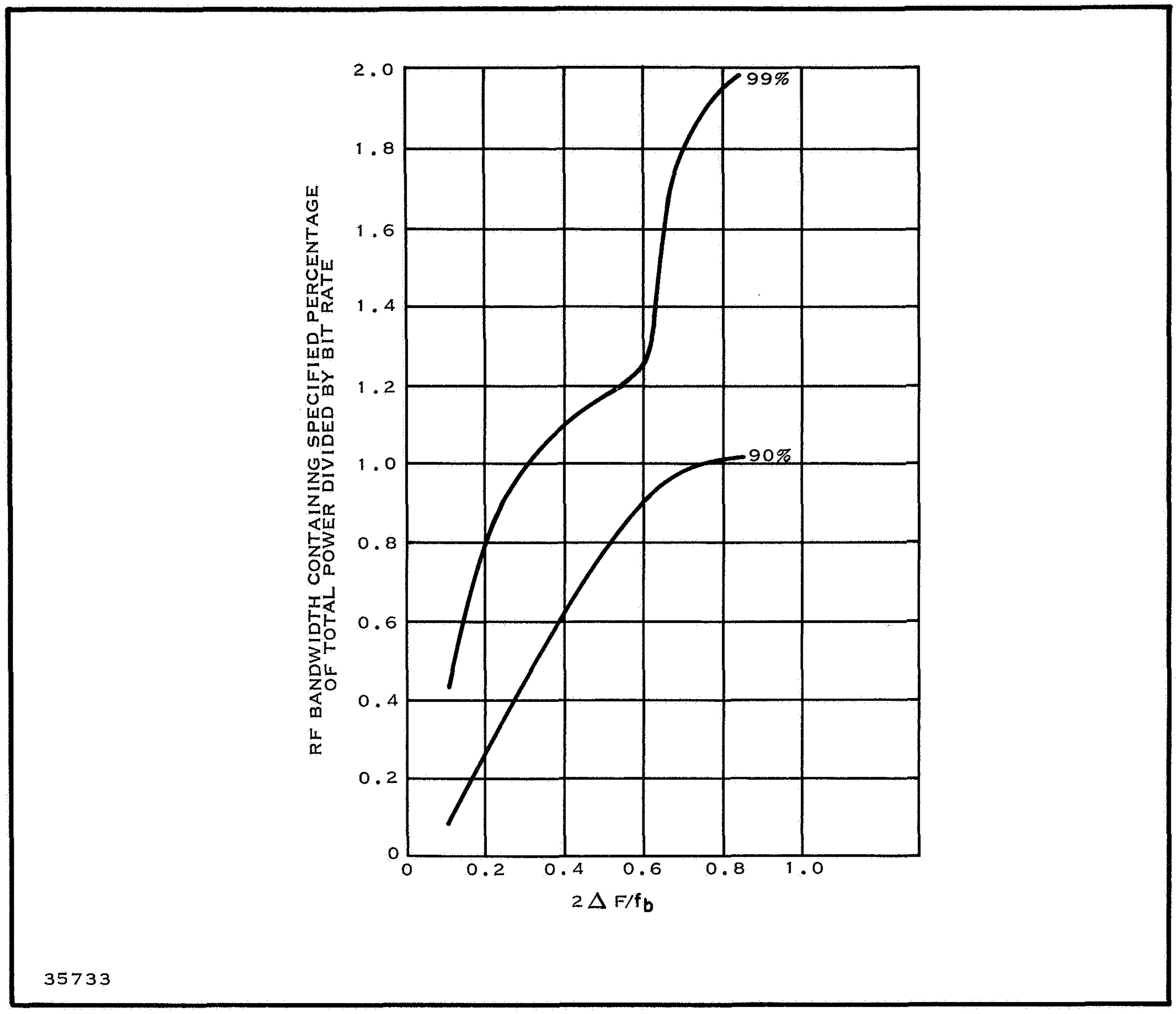

Figure 2. Band Occupancy of Carrier Frequency Modulated With Random Binary Sequence of Bit Rate f_b With Peak-to-peak Deviation $2\Delta F$ (No Premodulation Filter)

We may now compare the bandwidth requirement for PCM/FM with the requirement for linear FM previously established. For signal circuits, we require that the bandwidth be large enough to contain 99 percent of the power in the modulated wave. From Table III, we note that the maximum bandwidth condition occurs for a deviation ratio of 0. 7 with no premodulation filtering; this bandwidth is 7. 67 MHz. For the linear case, 8. 0 MHz is needed and establishes the requirement. For AFC circuitry, it is sufficient that 90 percent of the signal spectrum be passed. From Table III, the largest bandwidth also occurs for a deviation ratio of 0. 7 with no filtering; this bandwidth is 4. 16 MHz.

For the linear FM case, this requirement is 5.2 MHz and is also the larger of the two values. Accordingly, the bandwidth specifications are 8.0 MHz for signal circuits and 5.2 MHz for AFC circuits.

B. INDIRECT FREQUENCY MODULATION

Basic FM transmitter configurations may be classified according to the method of generating the frequency modulation—direct or indirect—and, if direct, according to the method of AFC. The advantage of the indirect method of FM, which is accomplished by phase modulating an oscillator with the integral of the basic modulating signal, is that no AFC is required. This results from the fact that a crystal oscillator may be easily phase modulated. The disadvantage is the large amount of multiplication usually required following the phase-modulated oscillator in order to build up the required frequency deviation in the output modulated wave. Whether the indirect FM approach is feasible depends on the amount of multiplication required with the attendant circuit complexity in comparison with other approaches. The following analysis establishes the amount of multiplication required.

In frequency modulation, the modulating signal V(t) modulates the carrier $A_c \cos \omega_c t$ such that the resulting modulated wave is

$$M_f(t) = A_c \cos [\omega_c t + K_1 \int^t V(t)\, dt] \qquad (1)$$

where K_1 is a constant having the units of radians/second/volt. Similarly, the expression for a phase modulated wave is

$$M_p(t) = A_c \cos [\omega_c t + \phi(t)] \qquad (2)$$

where $\phi(t)$ is the instantaneous departure in the phase of the modulated wave from the phase of the unmodulated wave. Thus, if frequency modulation is to be accomplished with a phase-modulated transmitter, $\phi(t)$ must be [from equations (1) and (2)]

$$\phi(t) = K_1 \int^t V(t)\, dt \qquad (3)$$

This integration is performed by a circuit before the modulating signal reaches the modulator. For the linear FM case, the modulation may be considered to be a simple sinusoid of frequency f_m.

$$V(t) = A_m \cos \omega_m t \qquad (4)$$

Substituting Equation (4) in Equation (3) and performing the simple integration yields

$$\phi(t) = \frac{K_1 A_m}{\omega_m} \sin \omega_m t \qquad (5)$$

Returning to Equation (1), we note that the instantaneous frequency ω_i is given by

$$\omega_i = \omega_c + K_1 V(t) \tag{6}$$

Substituting Equation (4) in Equation (6) yields

$$\omega_i = \omega_c + K_1 A_m \cos \omega_m t \tag{7}$$

If the amplitude A_m in Equation (4) corresponds to full modulation, from Equation (7) the result is

$$K_1 A_m = 2\pi \Delta F \tag{8}$$

where ΔF is the peak frequency deviation. Substituting Equation (8) in Equation (5) yields

$$\phi(t) = \frac{\Delta F}{f_m} \sin \omega_m t \tag{9}$$

from which the peak phase deviation is obviously

$$\Phi = \frac{\Delta F}{f_m} \tag{10}$$

The maximum value of Equation (10) occurs for the lowest modulation frequency to be encountered, $f_m = f_\ell$. For this specific case, with $\Delta F = 1.5$ MHz and $f_\ell = 2.0$ KHz, the peak phase deviation Φ is 750 radians.

Only a small amount of phase deviation can be obtained in the modulator if the distortion is to be held to a low value; a typical value is 0.3 radian. Thus, a multiplication of 750/0.3 = 2500 is required in order to achieve the desired frequency deviation in the output signal. When the multiplication factor is this high, frequency translation (down conversion) is necessary and is typical of the indirect method of frequency modulation. A block diagram of this type of transmitter is shown in Figure 3.

The large amount of multiplication necessary with attendant spurious signals forces the use of stable, narrow bandpass filters as interstage elements or bandpass filters in combination with a phase-locked loop. An additional problem with this system is associated with the integrator. This circuit must function as an integrator over a frequency range of 2.0 KHz to 1.0 MHz, a range of nearly nine octaves. The indirect method of frequency modulation is particularly unsuited to implementation in integrated circuitry because of the relatively large number of multiplier stages and narrow bandpass filters needed. For these reasons, it has been rejected in favor of the direct FM approach.

C. DIRECT FREQUENCY MODULATION

A variety of basic configurations of direct frequency modulation transmitters are discussed in Section IV. Almost all of these are alike in VCO, the driver stages, and the power amplifier. They differ in the method of achieving AFC. Even so, the basic requirements of open-loop gain and the

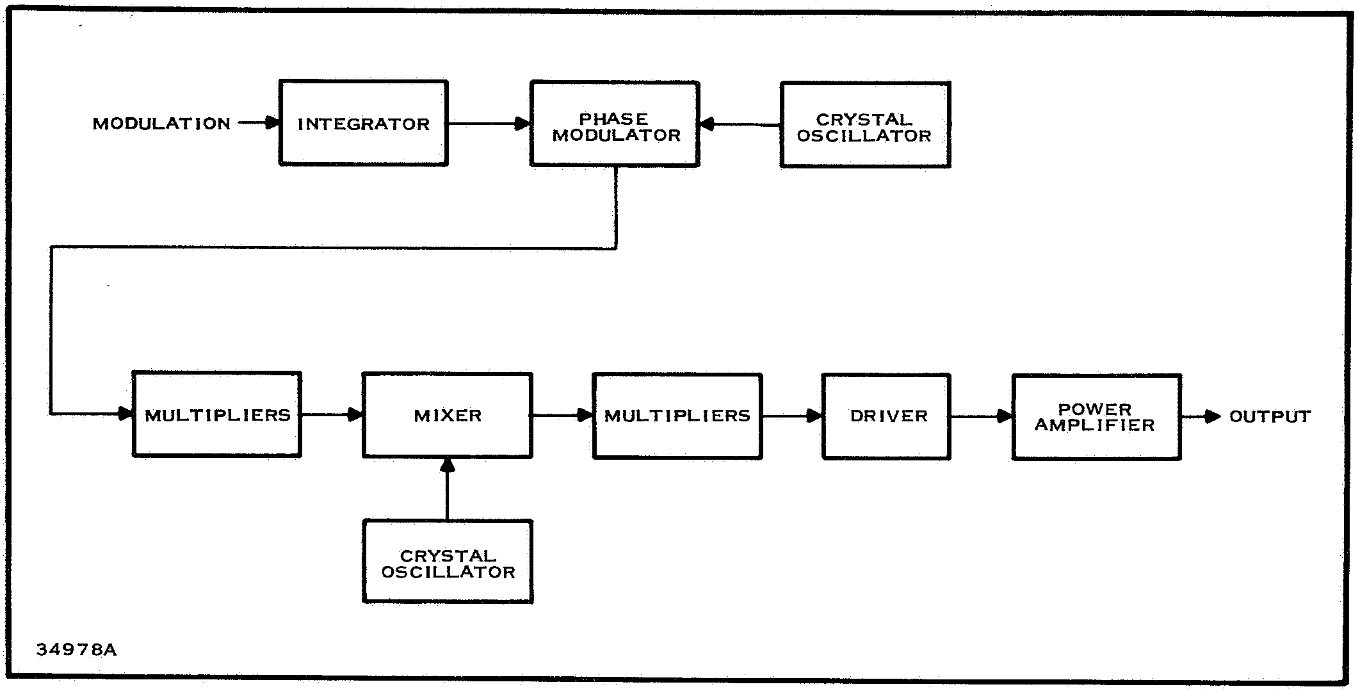

Figure 3. FM Transmitter—Indirect Method of Modulation

cutoff frequency of the low-pass filter through which the error signal passes before applying compensation to the VCO are essentially the same. The open-loop gain determines the fractional correction which can be applied to the uncompensated VCO center frequency. The open-loop gain and the cutoff frequency of the low-pass filter determine the characteristics of the system in its speed of response and amount of modulation distortion. On the one hand, we desire a fast response to minimize the effect of a changing environment, but on the other hand, we need a relatively slow response to avoid distortion of the modulation. The general requirements are reviewed here and specific requirements for particular AFC systems are reviewed in the following section.

1. Open-loop Gain

All the AFC systems function essentially as shown in Figure 4A. The instantaneous output frequency is compared with a stable reference frequency and an error voltage derived. This voltage is passed through a low-pass filter to remove modulation components and then applied to the VCO to reduce the error in the center frequency of the VCO. The error voltage may be obtained from a discriminator operating with a mixer and the multiplied output of a crystal oscillator, or from a discriminator with the input switched between the signal and the reference, or from some form of pulse discriminator. The result is basically the same in all cases. Figure 4B shows this basic loop with the summing junctions and transfer functions of the operating elements. (ω's denote the time domain quantities

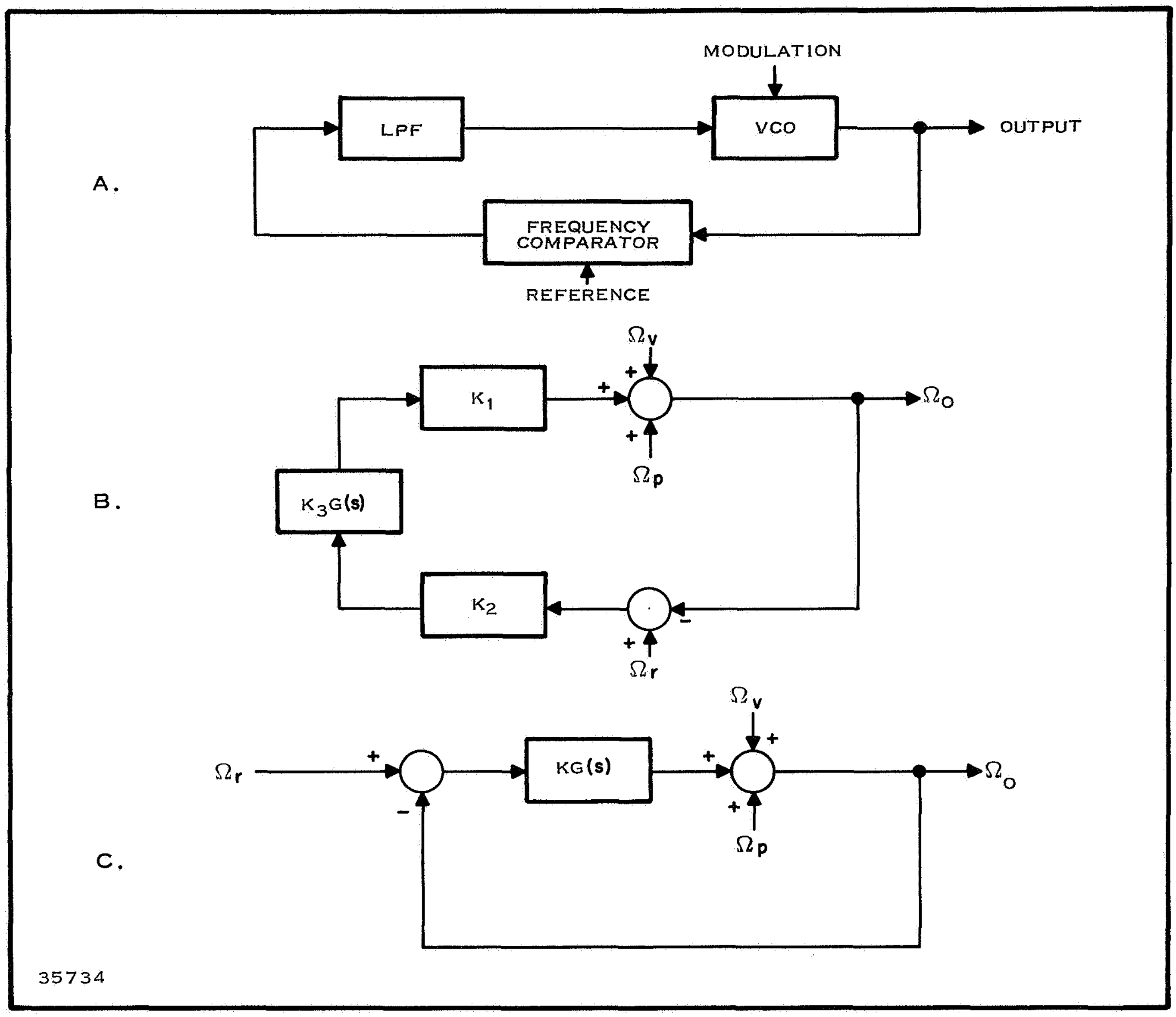

Figure 4. Basic AFC Loop

and Ω's the s-domain quantities.) The output frequency ω_o is the sum of the VCO uncompensated center frequency, ω_v; the open-loop component of the instantaneous frequency due to the modulation, ω_p; and a compensation term obtained by comparing the output of the VCO with the reference frequency, ω_r. Also, K_1 is the VCO constant expressed in radians/second/volt, and K_2 is the discriminator constant expressed in volts/radian/second; K_3 is dimensionless and includes the gain (loss) of amplifiers, frequency converters, the filter and summing junctions. The term G(s) includes the frequency variant part of the open-loop transfer function, the most significant part of which is the low-pass filter transfer function. In Figure 4C the various gains have been combined into a single dimensionless gain constant, K.

The closed-loop response for the simple system of Figure 4C is

$$\Omega_o = \frac{\Omega_v + \Omega_p + KG(s)\Omega_r}{1 + KG(s)} \tag{11}$$

If the low-pass filter is a single section RC network and the 3-dB cutoff frequency is very much lower than other circuit elements in the system, we have

$$G(s) = \frac{1/RC}{S + (1/RC)} \tag{12}$$

Substituting Equation (12) in Equation (11) (noting that ω_v and ω_r are constants and employing the final value theorem), we obtain the steady-state solution of Equation (11), with no modulation:

$$\omega_o = \frac{\omega_v}{1 + K} + \frac{K\omega_r}{1 + K} \tag{13}$$

Equation (13) may be arranged to yield

$$\omega_o = \omega_r + \frac{(\omega_v - \omega_r)}{1 + K} \tag{14}$$

Equation (14) shows that the output frequency differs from the reference frequency ω_r by an amount equal to the deviation of the uncompensated VCO frequency from the reference divided by the open-loop gain (for $K >> 1$). Unfortunately, the reference frequency itself differs from the assigned output frequency. The combined effect of the stabilities of the two oscillators, VCO and reference, produces an output stability [from Equation (14)] of

$$\delta_o = \delta_r + \frac{\delta_v - \delta_r}{1 + K} \tag{15}$$

All frequencies are assumed to be approximately the same, where δ_o is the output frequency stability in percent, δ_v the VCO frequency stability in percent, and δ_r the reference oscillator stability in percent. It is, of course, evident that the worst case occurs when δ_r and δ_v have the same sign. Rearrangement of Equation (15) yields

$$K = \frac{\delta_v - \delta_o}{\delta_o - \delta_r} \tag{16}$$

A value of $\delta_v = 2$ percent should be allowed to cover all changes in the VCO frequency, including initial manufacturing tolerance, variations due to temperature as well as aging effects. Since the ultimate objective is to provide an extremely small transmitter, the crystal used will probably have a combined tolerance and long-term drift of ±0.002 percent. The combined effects of these two oscillators on the output frequency determine the value of K required to meet the output frequency stability specification of 0.005 percent. Substitution of these values in Equation (16) shows that the required value of K is 667, or 56.5 dB. Equation (16) is plotted in Figure 5 for a range of parameter values normally of interest. The main thing to note

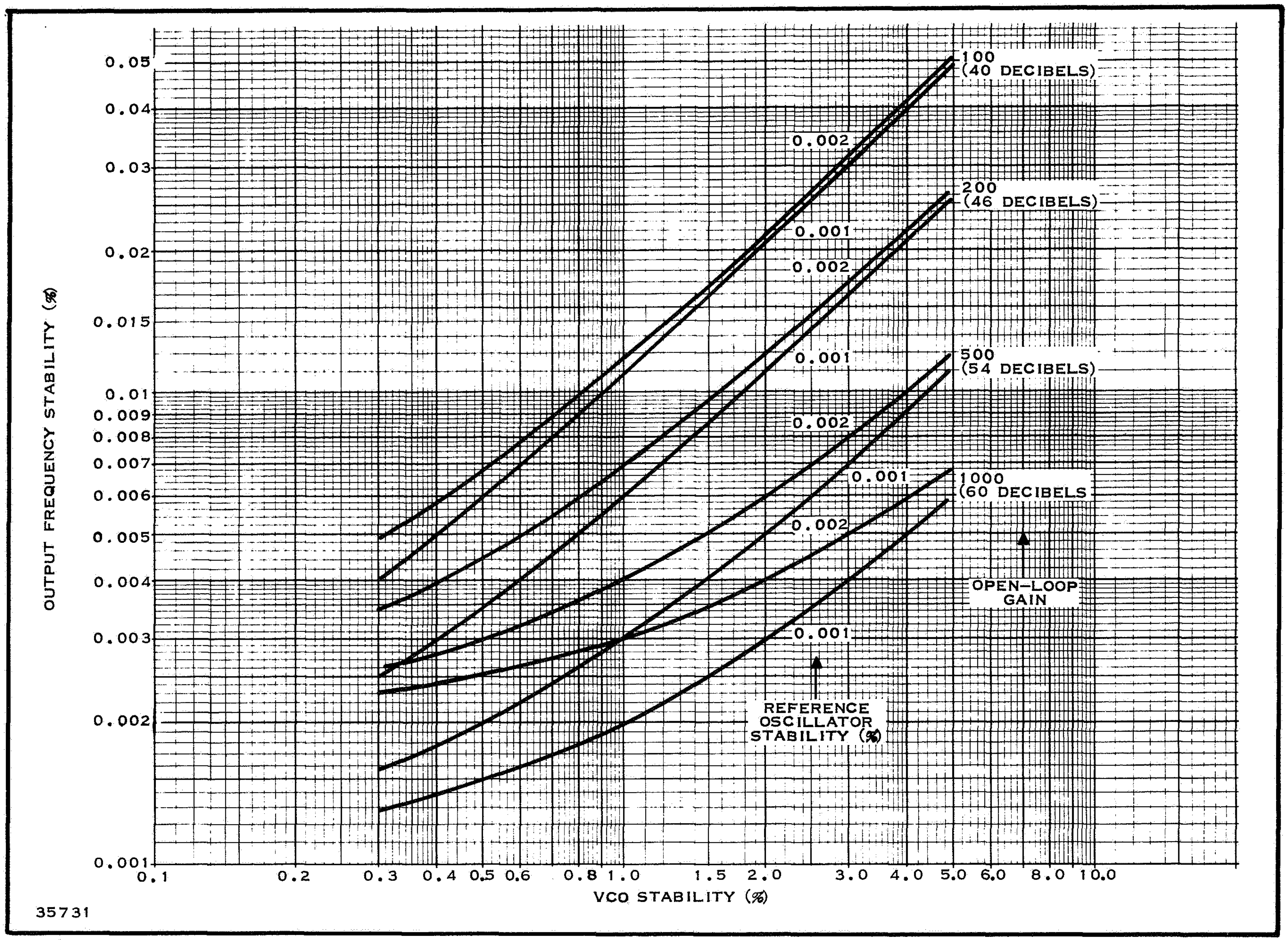

Figure 5. Output Frequency Stability as a Function of VCO Stability, Open-loop Gain, and Reference Oscillator Stability

from this figure is that there is little point in improving the stability of the referenced oscillator if the VCO stability is poor and the open-loop gain is low. However, as VCO stability is improved and open-loop gain increased, the output frequency stability becomes more dependent on the reference oscillator stability. Under these conditions it becomes desirable to improve the reference oscillator stability.

2. Low-pass Filter Cutoff Frequency

Unfortunately, the closed-loop AFC system introduces distortion into the modulation component of the output. Returning to Equation (11), we note that the closed-loop transfer function for the modulation component is

$$\frac{\Omega_{op}}{\Omega_p} = \frac{1}{1 + KG(s)} \tag{17}$$

where Ω_p is the transform of the open-loop component of the instantaneous frequency due to modulation and Ω_{op} is the transform of the closed-loop component of the instantaneous frequency due to modulation. With the applied modulation a simple sinusoid, we obtain

$$\omega_p = K_1 V(t) = \Delta\omega \sin \omega_m t \tag{18}$$

where $\Delta\omega$ is the peak deviation due to the modulating sine wave of frequency ω_m. Substituting Equation (12) and the transform of Equation (18) into Equation (17) and taking the inverse transform, we obtain the steady-state component of the instantaneous frequency due to modulation:

$$\omega_{ops} = \Delta\omega \left\{ \frac{(\omega_3)^2 + (\omega_m)^2}{\left[(1 + K)\omega_3\right]^2 + (\omega_m)^2} \right\}^{1/2} \sin(\omega_m t + \beta) \tag{19}$$

$$\beta = \tan^{-1} \frac{\omega_m}{\omega_3} - \tan^{-1} \frac{\omega_m}{(1 + K)\omega_3} \tag{20}$$

where $\omega_3 = 1/RC$, the 3-dB cutoff frequency for the low-pass filter, expressed in rad/s. When Equation (19) is compared with Equation (18), we see that frequency-dependent amplitude and phase terms have been introduced. These represent distortion of the desired modulation as given by Equation (18).

The amount of amplitude distortion given by the ratio of the amplitudes of Equations (19) and (18) is

$$\alpha = 10 \log_{10} \left[\frac{(1 + K)^2 \omega_3{}^2 + \omega_m{}^2}{\omega_3{}^2 + \omega_m{}^2} \right] \text{(attenuation in dB)} \tag{21}$$

Inspection of Equation (21) shows that α approaches 0 dB as the modulation frequency is increased. Thus, the effect of the closed-loop AFC system is to introduce a droop in the response at the low-frequency end of the modulation baseband. Equation (21) is plotted in Figure 6 for three values of attenuation. For our specific case, we assume that a 0.1-dB droop in the response at the lowest modulation frequency, 2 KHz, can be allowed. With

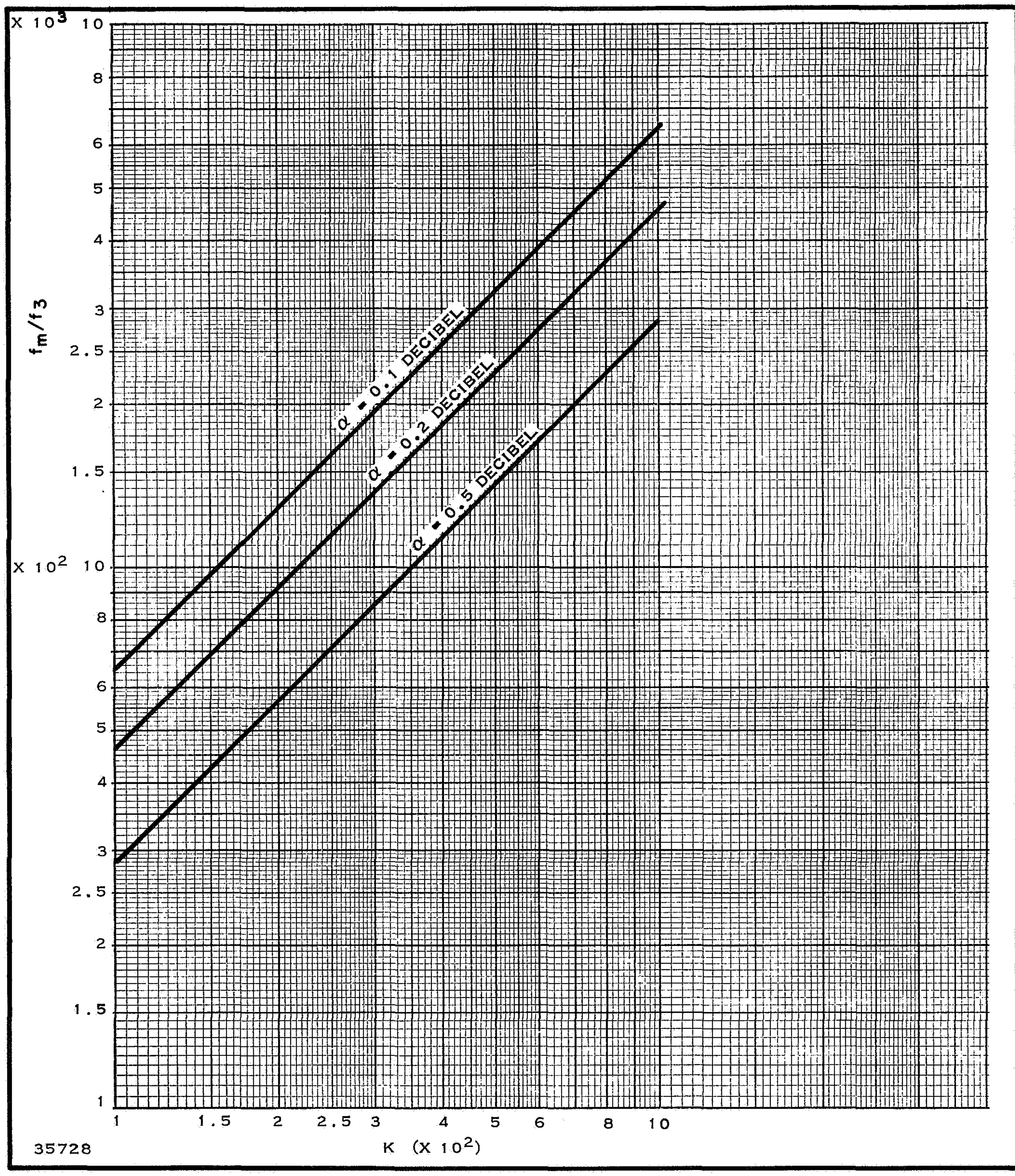

Figure 6. Ratio of the Modulation Frequency f_m to the 3-dB Cutoff Frequency of the Low-pass Filter f_3 as a Function of the Open-loop Gain K for Three Values of Attenuation of the Amplitude Response at f_m

the loop gain previously set at 667, the 3-dB cutoff frequency of the low-pass filter is [from Equation (21)] 0.46 Hz. This is not the limiting case, as will be seen.

The phase shift introduced into the modulation by the closed-loop AFC system is given by Equation (20). This phase function must be either negligibly small for all frequencies across the baseband or it must be a linearly increasing function of frequency. Any departure from linearity introduces time-delay distortion. Furthermore, the phase angle must be zero for zero frequency, if zero intercept distortion is to be avoided. This zero-intercept criterion may be ignored for a baseband consisting of modulated subcarriers; it cannot be ignored when the baseband consists of an unmodulated signal such as video.

Inspection of Equation (20) shows that the phase angle is zero at zero frequency, but it also approaches zero for high modulation frequencies. Using a series expansion for the arctangent function, we may rewrite Equation (20) as

$$\beta \approx \frac{K\omega_3}{\omega_m} - \frac{1}{3}\left(\frac{K\omega_3}{\omega_m}\right)^3 \text{ rad} \tag{22}$$

where $\omega_m/\omega_3 >> 1$, $K >> 1$, and $\omega_m/K\omega_3 > 1$. In general, the contribution of the second term on the right-hand side of Equation (22) is quite small; higher-order terms have been neglected. Since β will be essentially zero for the highest modulation frequency of interest, we would prefer that it be zero for all frequencies. Investigation of the derivative of Equation (20) shows that the maximum phase shift occurs when

$$\frac{\omega_m}{\omega_3} = \sqrt{1 + K} \tag{23}$$

Substitution of Equation (23) into Equation (20) gives the maximum phase shift:

$$\beta_{max} = \tan^{-1}\sqrt{1 + K} - \tan^{-1}\frac{1}{\sqrt{1 + K}} \tag{24}$$

Even for small values of K, such as $K = 100$, $\beta_{max} = 78.5$ degrees. Such a large phase shift cannot be tolerated. The maximum phase shift occurs at a value of ω_m/ω_3 much lower than we can allow. Thus, for our purposes the relative maximum value of β will occur at the lowest modulation frequency of interest.

To minimize the effect of time delay distortion, the following criterion is used: the variation of phase shift across the baseband from the best straight-line fit cannot exceed the equivalent of $\pm\pi/2$ radians phase shift at the highest modulation frequency. Unfortunately, the greatest departure from the best straight-line fit—which is actually zero phase shift for all frequencies—occurs at the lowest frequency where the equivalent of $\pm\pi/2$ radians at the highest modulation frequency is small. The allowed value is

$$\beta_\ell = \frac{\pi}{2}\,\frac{\omega_\ell}{\omega_h} \tag{25}$$

where β_ℓ is the maximum value of phase shift allowed at the lowest modulation frequency, and ω_ℓ and ω_h are the lowest and the highest modulation frequencies in the baseband, respectively. Investigation shows that the second term of Equation (22) may be neglected. Substituting Equation (25) into Equation (22) and ω_ℓ for ω_m in Equation (22), we obtain

$$\frac{\omega_\ell}{\omega_3} = \frac{2K}{\pi}\left(\frac{\omega_h}{\omega_\ell}\right) \tag{26}$$

This equation is plotted in Figure 7 for several values of the open-loop gain K. For the specific case of interest, K = 667, f_ℓ = 2 KHz, f_h = 1 MHz, and the 3-dB cutoff frequency for the low-pass filter is 0.00942 Hz, approximately 0.01 Hz. If the cutoff frequency is made higher than this value, excessive time delay distortion will result.

Depending upon the exact structure of the video baseband signal, it may be possible to increase the 3-dB cutoff frequency. If this baseband signal is structured similar to a conventional television signal, it is only necessary that the criterion discussed above be met for frequency components at the line rate and above. For lower frequency components, only the step response is important. Too much droop in the step response for a period equal to the lowest modulation frequency involved produces shading down of the reproduced picture.

Using the transfer function of Equation (17) and the same single section low-pass filter whose transfer function is given by Equation (12), the step response can be shown to be

$$\omega_{opt} = \frac{1 + K \exp\left[-(1 + K)\omega_3 t\right]}{1 + K} \tag{27}$$

Since the lowest modulation frequency is 2.0 KHz, we are concerned with the amount of droop in the step response after a time of 250 μs, one-half period at 2.0 KHz square wave. Based on Equation (27), the required 3-dB cutoff frequency as a function of the open-loop gain K for various amounts of droop in the step response has been determined (Figure 8).

In general, 5- to 6-percent droop can be tolerated. This is the equivalent of 0.5 dB. With the loop gain set at 667, a low-pass cutoff frequency of 0.055 Hz produces 0.5 dB droop.

Table IV summarizes the results of this analysis of the requirement for the low-pass filter. Clearly, time-delay distortion considerations establish the requirement for a 0.01-Hz cutoff frequency.

Table IV. Low-pass Filter 3-dB Cutoff Frequency Requirement for Three Different Modulating Signal Characteristics

Characteristic	Specification	Required 3-dB cutoff frequency for low-pass filter
Response to a sine wave at the lowest modulating frequency, 2.0 KHz	0.1 dB down	0.46 Hz
Time-delay distortion across total baseband	250 ns	0.01 Hz
Step response	0.5 dB down 250 ms after step	0.05 Hz

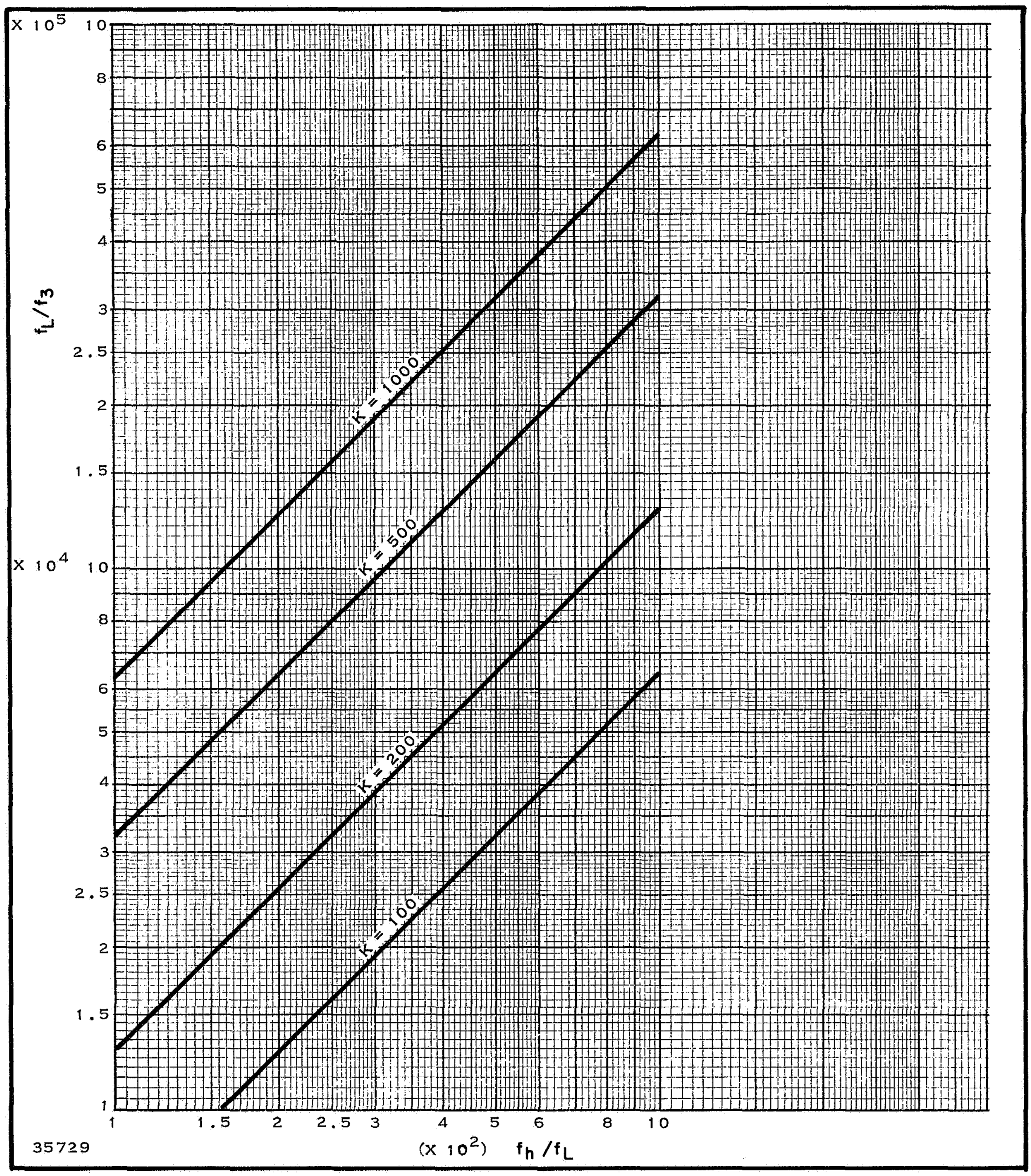

Figure 7. Ratio of the Lowest Modulation Frequency f_ℓ to the 3-dB Cutoff Frequency of the Low-pass Filter f_3 as a Function of the Ratio of the Highest Modulation Frequency f_h to the Lowest Modulation Frequency for Four Values of the Open-loop Gain K (Allowed Time-delay Distortion: One-quarter Period at the Highest Modulation Frequency)

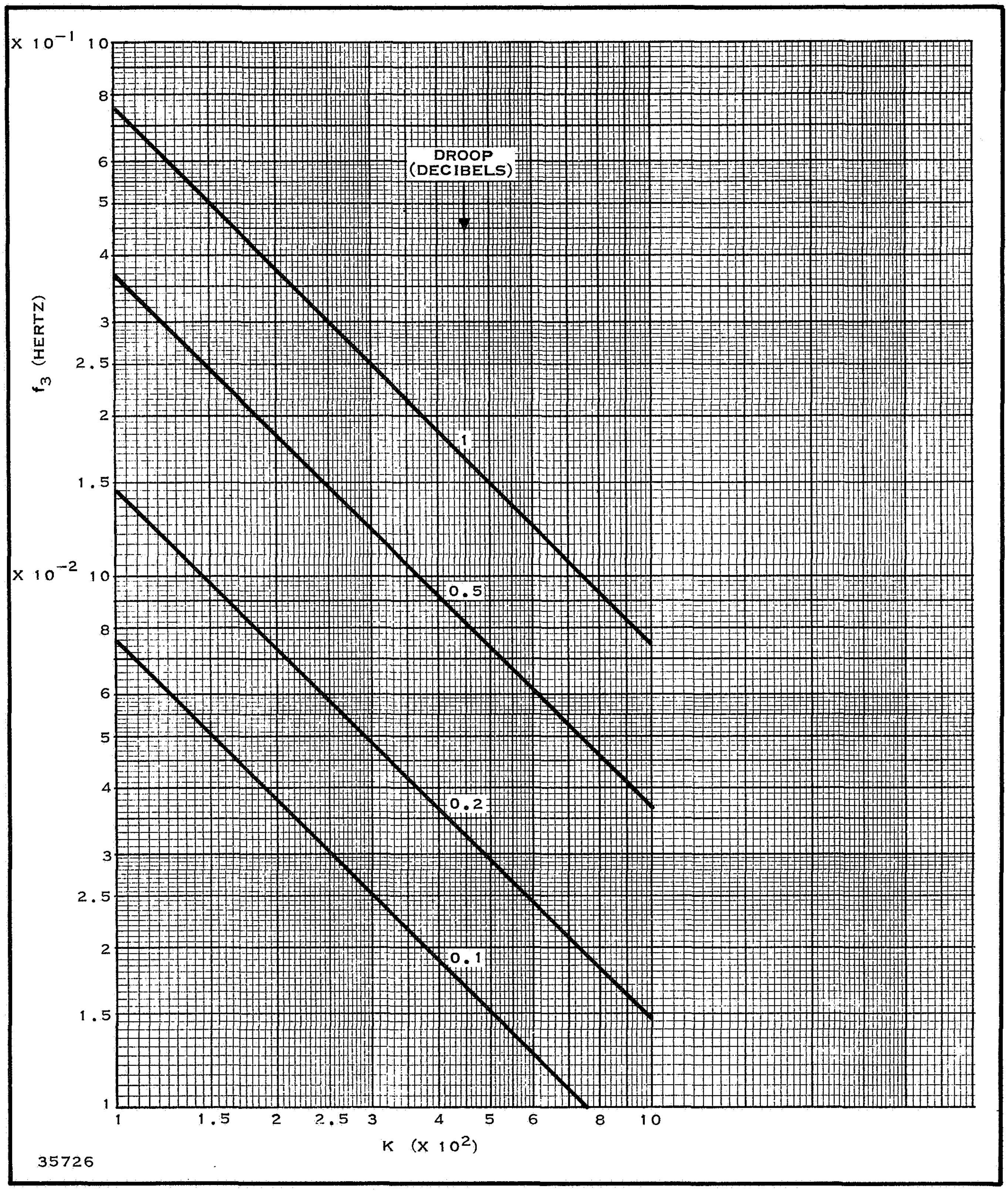

Figure 8. 3-dB Cutoff Frequency of the Low-pass Filter as a Function of Open-loop Gain K for Four Degrees of Droop at the End of a 250-μs Period (One-half Period at 2.0 KHz)

SECTION IV

DIRECT FM TRANSMITTER CONFIGURATIONS

To achieve the center frequency stability required, 0.005 percent, some form of crystal control is essential. There are three basic approaches to realizing this frequency control in direct FM telemetry transmitters.

The simplest approach uses a voltage-controlled crystal oscillator (VCXO) that is capable of providing the center frequency stability while allowing frequency modulation. This system, however, has fundamental limitations; these will be discussed.

In the second basic approach, the objective is to allow the use of a relatively unstable voltage-controlled oscillator (VCO) but also attempting to "swamp out" the instability in the output frequency by employing a crystal-controlled oscillator whose frequency is mixed with the VCO frequency. The sum of the two frequencies establishes the output frequency. The larger the ratio of the crystal-oscillator frequency to the VCO frequency, the more dependent the output frequency stability becomes on the crystal-oscillator stability and the less dependent it becomes on the VCO stability. This system has its limitations, as will be seen.

The third approach uses an unstable VCO operated with an automatic-frequency-control (AFC) feedback loop. Of the three approaches, this is the only one capable of meeting the performance requirements for this transmitter. Nine variations of the AFC system are considered in this section.

A. DISCRIMINATOR AFC

One of the oldest and simplest techniques used for AFC is shown in Figure 9. The VCO output and the reference oscillator output are mixed to derive a difference frequency. This difference frequency is then compared with the center frequency of the discriminator, and an error voltage is developed and fed back to the VCO. The feedback is negative and the circuit functions to reduce the discrepancy between the difference frequency and the discriminator center frequency. The arrangement is simple and effective, within limits. The limitation is due to the dependence of the controlled variable on the tolerance and stability of the discriminator center frequency.

The reference signal developed by the crystal-controlled oscillator can be made to fall within ±0.001 to ±0.002 percent of the desired frequency without oven control, including the combined effects of the initial manufacturing tolerance, long-term drift, and short-term instability. This is two orders of magnitude less than the combined tolerance and stability effects associated with the discriminator center frequency. (Crystal discriminators are not applicable because of their narrow bandwidth.)

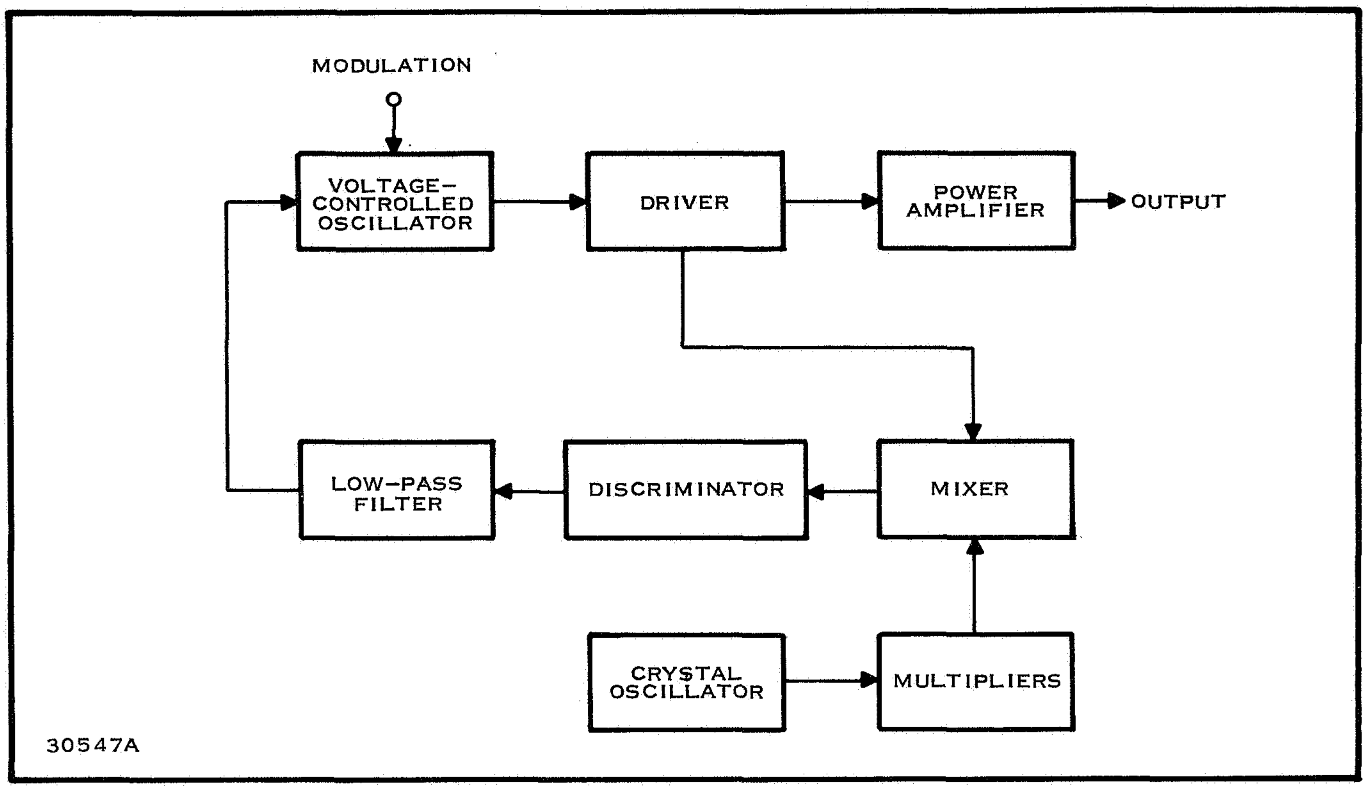

Figure 9. Discriminator AFC—FM Transmitter

For this reason, it is essential that the influence of the discriminator on the average frequency of the VCO be minimized. This may be accomplished simply by choosing the reference frequency close to the unmodulated output frequency of the VCO. This makes the difference frequency small and the center frequency of the discriminator low, thus reducing the effect of inaccuracies in the center frequency of the discriminator on the output frequency of the VCO. This situation may be investigated with the aid of Figure 10.

In Figure 10A, V(s) represents the modulating voltage applied to the VCO; K_1 is the VCO constant expressed in radians/second/volt input, and Ω_v is the uncompensated center frequency of the VCO. The output frequency Ω_o is subtracted from the reference frequency Ω_r in the mixer and added to the center frequency of the discriminator Ω_d. The polarities shown are, of course, for negative feedback and assume low side injection; for example, the reference oscillator frequency is less than the output frequency. The discriminator conversion constant (shown as the block labeled K_2) is expressed in volts/radian/second input. The transfer function $K_3G(s)$ represents the low-pass filter characteristic. The block diagram of Figure 10B is a simplification of the diagram of Figure 10A. The open-loop gain K is the product of K_1, K_2, and K_3. The modulation has been referred to the output and is represented as Ω_p; this is the transform of the instantaneous frequency component of the output due to the modulation.

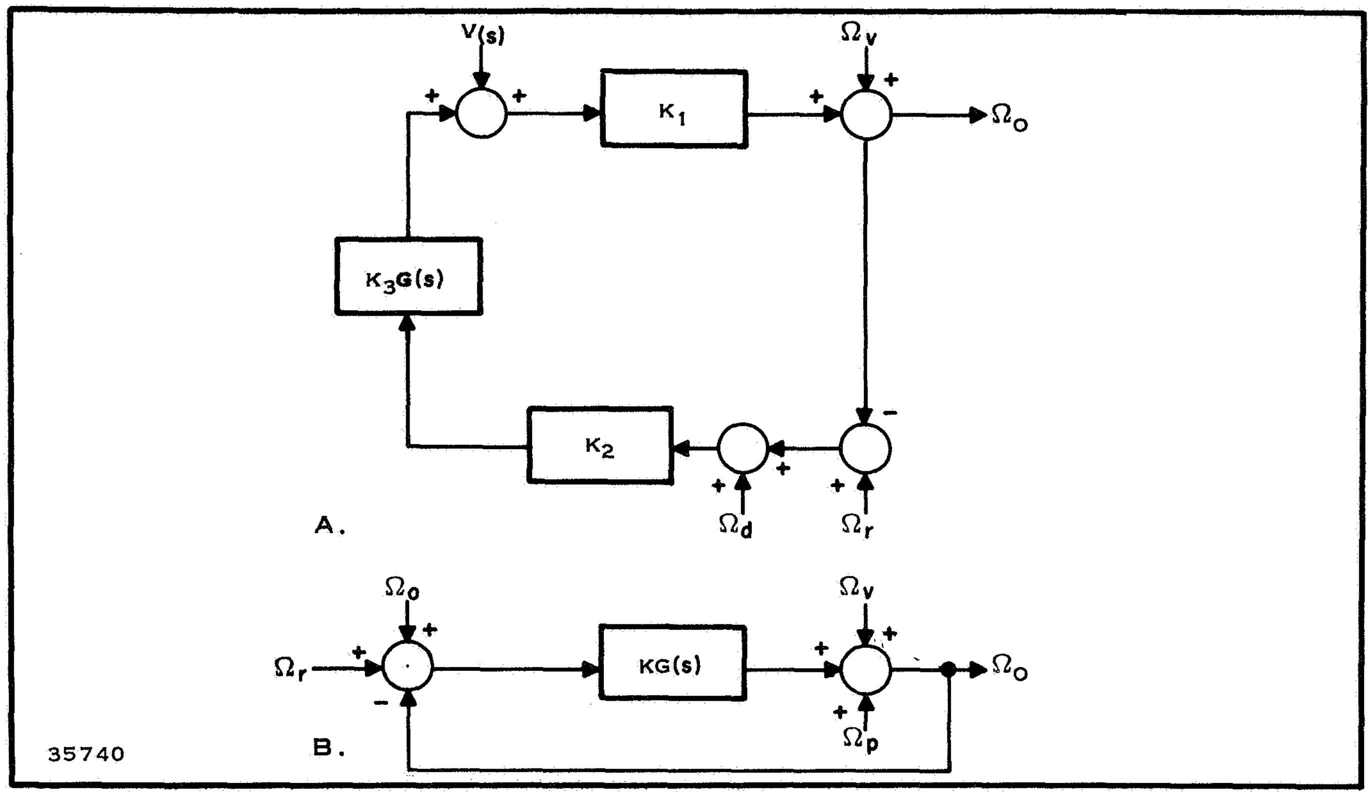

Figure 10. Discriminator AFC Loop

The closed-loop response for the system of Figure 10B is

$$\Omega_o = \frac{\Omega_v + \Omega_p + KG(s)\ (\Omega_d + \Omega_r)}{1 + KG(s)} \tag{28}$$

With a single section low-pass RC filter, G(s) is given by

$$G(s) = \frac{1/RC}{S + (1/RC)} \tag{29}$$

Substituting Equation (29) in Equation (28) and assuming no modulation, the steady-state response is given by

$$\omega_o = \frac{\omega_v + K(\omega_d + \omega_r)}{1 + K} \tag{30}$$

where ω_v, ω_d, and ω_r are all constants. Equation (30) may be rearranged to yield

$$\omega_o = \omega_r + \omega_d + \frac{\omega_v - (\omega_r + \omega_d)}{1 + K} \tag{31}$$

The last term on the right-hand side of Equation (31) is the amount by which the output frequency differs from the desired value, which is the sum of the reference frequency and discriminator center frequency.

From Equation (30), the effect of the various center frequency stabilities may be determined with

$$\delta_o \omega_o' = \frac{\delta_v' \omega_v' + K(\delta_d \omega_d' + \delta_r \omega_r')}{1 + K} \tag{32}$$

The primed variables denote design center values, for example

$$\left.\begin{aligned} \omega_o' &= \omega_d' + \omega_r' \\ \omega_o' &= \omega_v' \end{aligned}\right\} \tag{33}$$

where

δ_o is the output frequency stability in percent
δ_v is the VCO center frequency in percent
δ_d is the discriminator center frequency in percent
δ_r is the reference oscillator center frequency in percent

Using the relations of Equation (33) in Equation (32), we can solve for the ratio of the design center values of the discriminator center frequency to the output frequency in terms of the various stability factors and the open-loop gain K. The result is

$$\frac{\omega_d'}{\omega_o'} = \frac{(1 + K)\,\delta_o - \delta_v - K\delta_r}{K(\delta_d - \delta_r)} \tag{34}$$

For the special case of $K \rightarrow \infty$

$$\frac{\omega_d'}{\omega_o'} = \frac{\delta_o - \delta_r}{\delta_d - \delta_r} \tag{35}$$

These two equations are plotted in Figure 11 as a function of the discriminator center frequency stability for a required output frequency stability of 0.005 percent and a reference oscillator stability of 0.001 percent.

The discriminator center frequency stability probably cannot be held closer than 0.25 percent including the combined effects of initial tolerance, ambient temperature changes, and long-term aging effects. Under these conditions and with infinite open-loop gain, the ratio of the discriminator center frequency to the RF output frequency is 0.016, or 1.6 percent. Unfortunately, the long-term stability of the VCO is expected to be in this same range, 1 to 2 percent. Thus, the combined requirement dictates the need for a discriminator having a bandwidth equal to approximately two times the center frequency. This approach is unsuitable for another reason: The

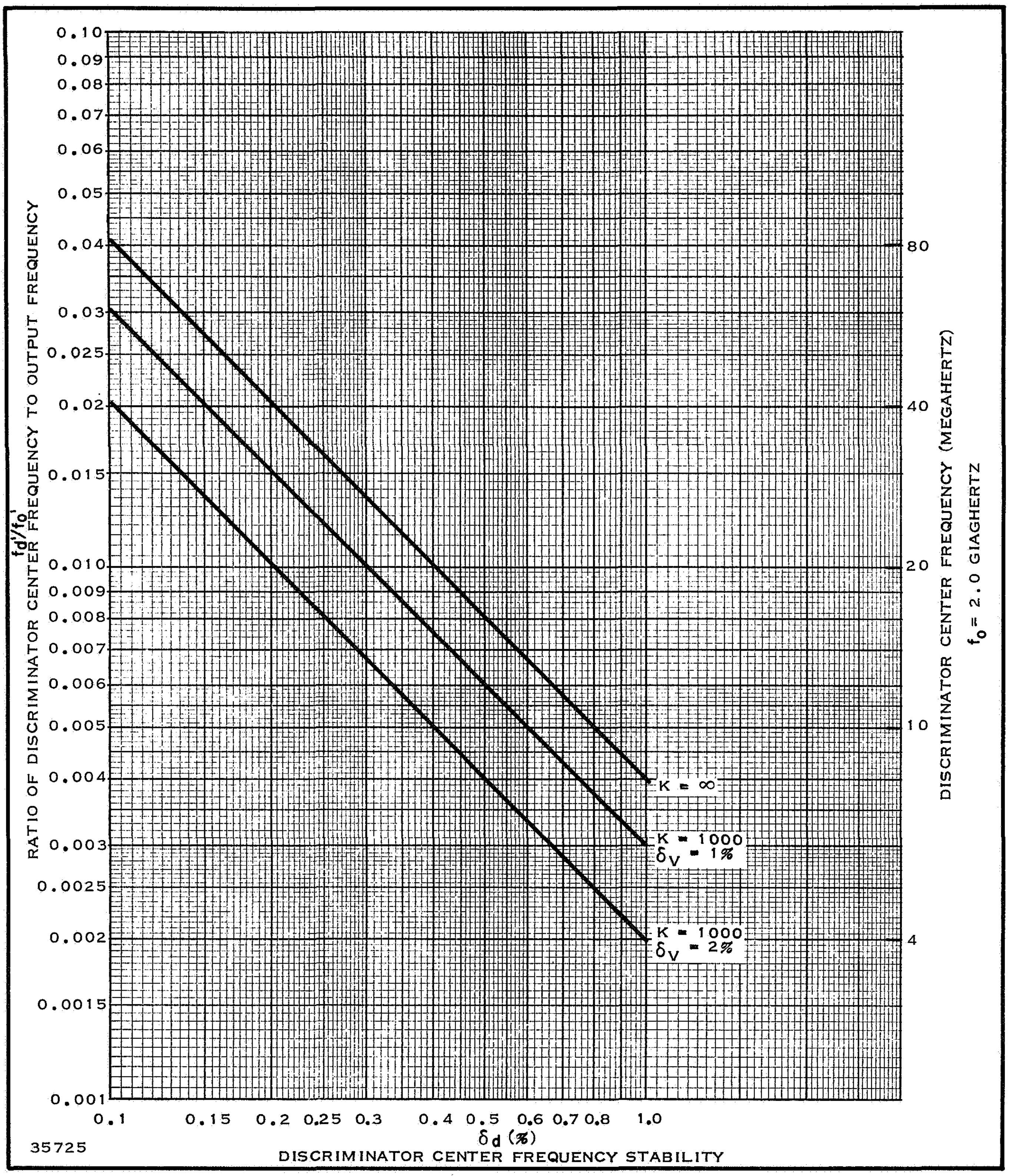

Figure 11. Ratio of the Discriminator Center Frequency to the RF Output Frequency as a Function of the Discriminator Center Frequency Stability for a Reference Oscillator Stability of 0.001 Percent and a Required Output Stability of 0.005 Percent

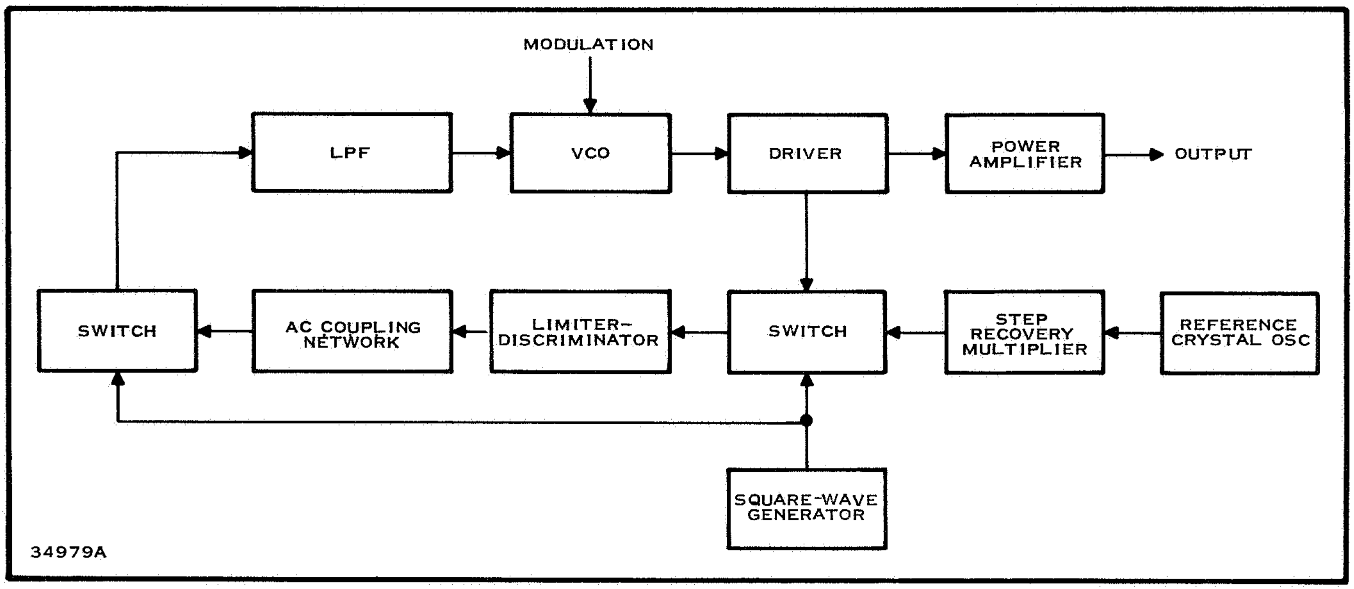

Figure 12. Gated Discriminator AFC—FM Transmitter

discriminator is far too low for effective implementation in integrated circuitry. For these reasons, this simple discriminator AFC was not considered further.

B. GATED DISCRIMINATOR AFC

The gated discriminator AFC system (Figure 12) is capable of performance that is significantly improved over that of the simple discriminator AFC system. In the improved system, the center frequency stability of the discriminator does not influence the stability of the output frequency. The square-wave generator of Figure 12 alternately switches either the RF output frequency or the reference frequency into the limiter-discriminator for equal periods of time. The reference frequency and the discriminator center frequency are chosen equal to the desired output frequency.

In the operation of the system, Figure 13, the discriminator center frequency f_d can be different from the reference frequency f_r (Figure 13A). The system functions to correct the discrepancy between the output frequency f_o and f_r by developing the associated voltages e_o and e_r at the output of the discriminator. Figure 13B shows the time waveform resulting from the square wave-switching. The modulation is superimposed on the mean value of the output frequency and thus appears in the discriminator output. This signal cannot be used directly as an error signal because of the difference between f_d and f_r. If, however, the signal is simply ac coupled the dc offset is removed (Figure 13C). Finally, the signal is synchronously demodulated by a second switch driven from the square-wave generator. The output of this switch (Figure 13D) is then passed through the low-pass filter and applied to the VCO. In the case shown, f_r is lower than f_o; had the reverse been true, the polarity of the error signal of Figure 13D would be negative.

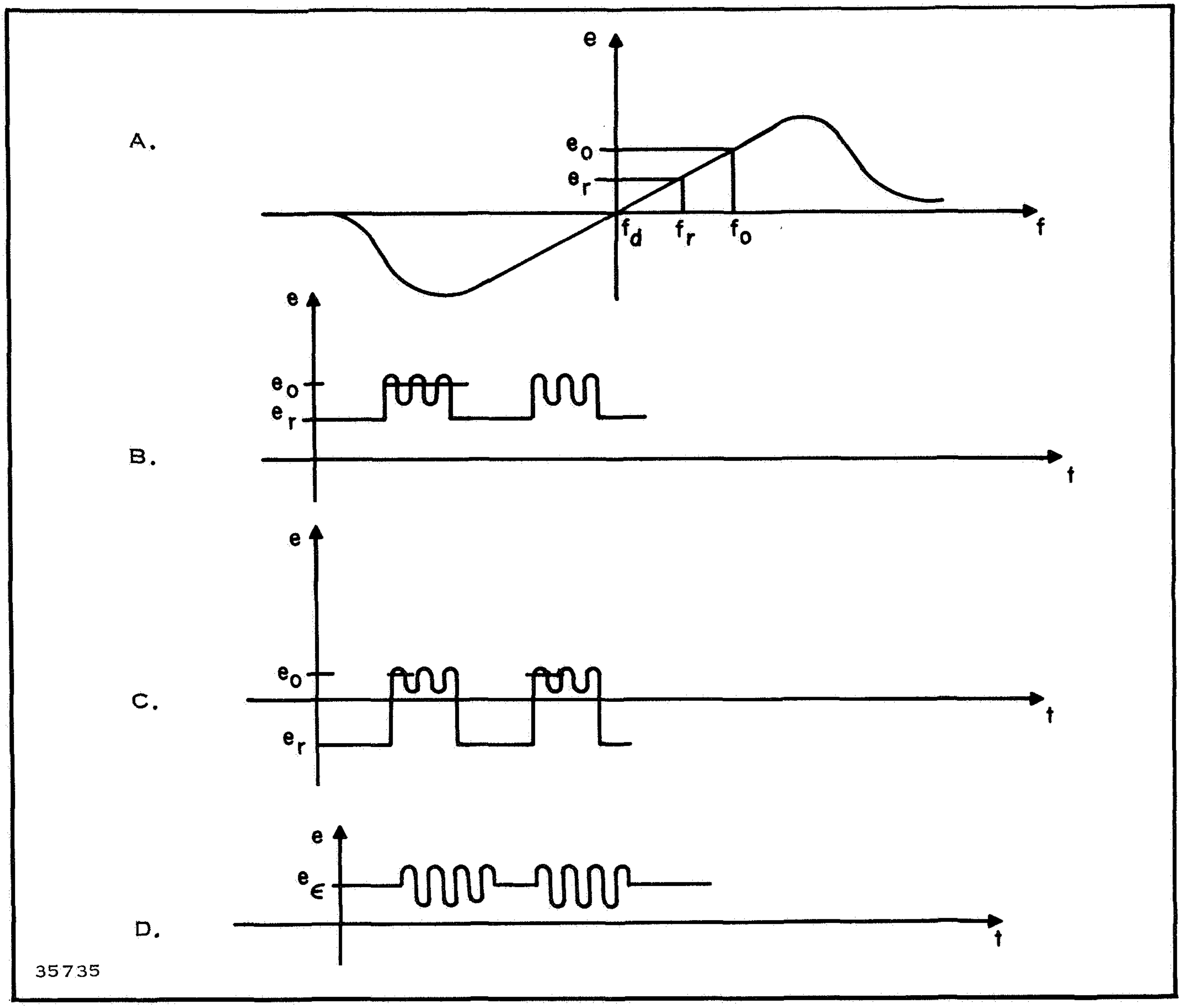

Figure 13. Gated Discriminator AFC Operation

This basic AFC system has been used in point-to-point microwave communications systems, in Telstar[5,6], and in telemetry transmitters[7]. However, the RF signals into the limiter-discriminator were at a relatively low frequency, around 70 MHz. Recent advances in PIN surface-oriented diodes have made it possible to switch at S-band; furthermore, these devices were designed for application to microwave integrated circuits.

The system functions exactly as described in Section III.C with regard to the open-loop gain requirement and low-pass filter characteristics. The system can be treated as a linear-feedback control system because the switching rate is much higher than the cutoff frequency of the low-pass filter. The choice of the switching frequency is, however, dependent on the modulating signal spectrum.

The low frequency content of the modulating signal spectrum along with the open-loop gain and the low-pass filter cutoff frequency place a constraint on the choice of the switching frequency. This can be seen by investigating the Fourier series of the periodic signal at the output of the discriminator. The square-wave switching function of unit amplitude can be shown to be

$$S(t) = \frac{1}{2}\left[1 + 2 \sum_{n=1}^{\infty} \frac{\sin n\frac{\pi}{2}}{n\frac{\pi}{2}} \cos n\omega_s t\right] \tag{36}$$

where ω_s is the square-wave switching frequency in rad/s, and n is the order of the harmonic. Based on the waveform of Figure 13D, which represents the output of the synchronous detector, and assuming the modulation to be a sinusoid of frequency f_m, the output of the discriminator is

$$e_\epsilon = e_{dc} + \frac{A_m}{2}\left\{\cos \omega_m t + \sum_{n=1}^{\infty} \frac{\sin n\frac{\pi}{2}}{n\frac{\pi}{2}} \left[\cos(\omega_m + n\omega_s)t + \cos(\omega_m - n\omega_s)t\right]\right\} \tag{37}$$

where

ω_m is the frequency of the modulating sinusoid in rad/s
e_{dc} is the dc value of the error voltage
A_m is the amplitude of the detected sinusoid.

The problem arises when $(\omega_m - n\omega_s)$ becomes small enough to pass through the low-pass filter and modulate the VCO. The amplitudes of the components of Equation (37) are shown in Figure 14. If any of the lower-order odd harmonics of the switching frequency fall near a significant low-frequency component of the modulation, very low-frequency modulation of the VCO can result.

Several things can be done to minimize or eliminate this effect. First, the switching frequency should be chosen such that none of its odd harmonics are near primary power frequencies or field rates of video signals. Second, a bandpass filter can be used at the output of the discriminator designed to pass the fundamental of the switching frequency and suppress the harmonics. Finally, the switching waveform can be designed to eliminate the harmonics; for instance, a sine-wave switching function could be used. This is not normally done since it places an added constraint on the matching of diodes used in the switch.

The extent of the problem can be determined by an analysis of the effect these spurious signals have on the modulated output of the transmitter. In Figure 15A the spurious signal ω_q is represented as being

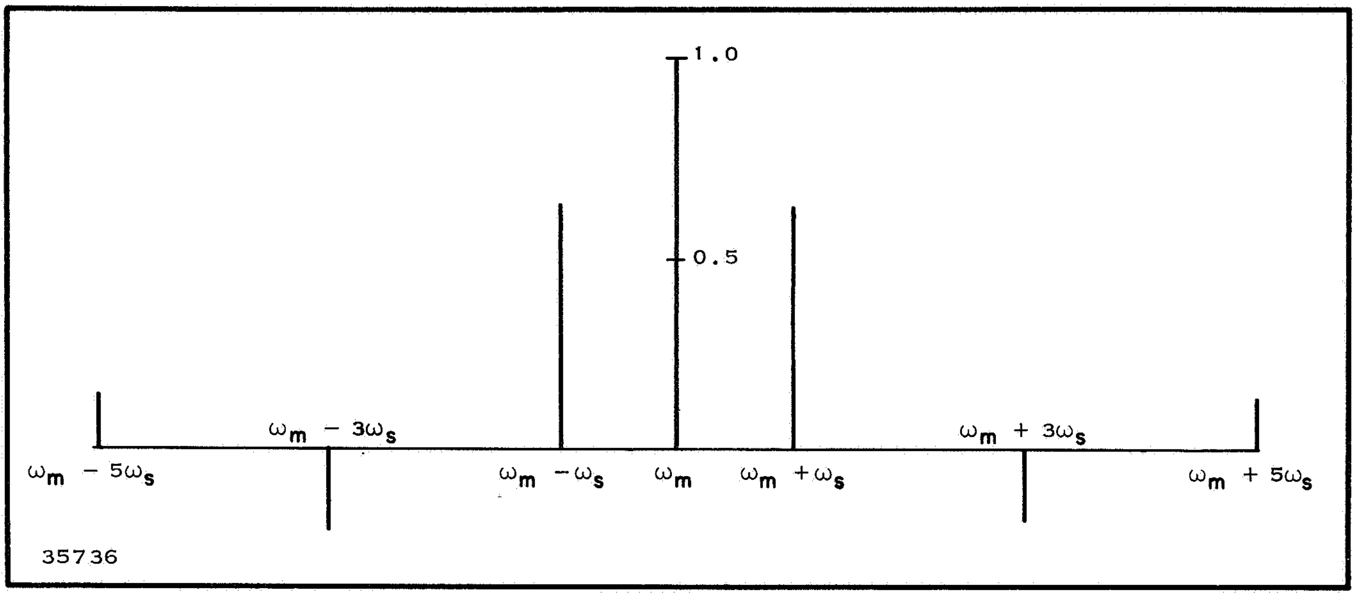

Figure 14. Relative Amplitudes of the Spectral Components Centered on the Modulation Frequency

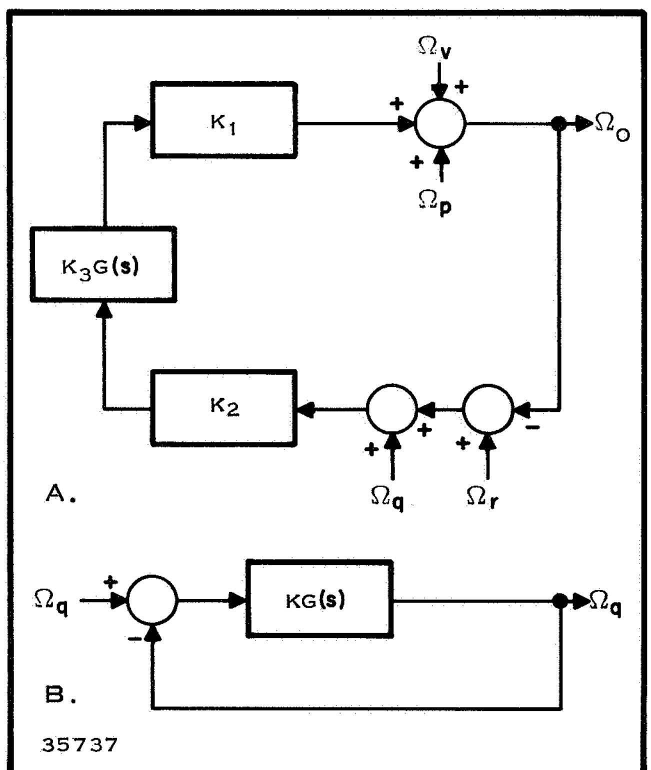

Figure 15. Gated AFC Loop for Study of Spurious Output

summed into the system at the input of the discriminator. The notation of Figure 15 is the same as that used in Figure 4. Thus, as far as the effect on the output is concerned, the simple diagram of Figure 15B may be used. Taking G(s) from Equation (12), we obtain the transfer function

$$\frac{\Omega_{oq}}{\Omega_q} = \frac{K\omega_3}{S + \omega_3\ (1 + K)} \qquad (38)$$

where the substitution $\omega_3 = 1/RC$ has been made and ω_3 is the 3-dB cutoff frequency of the low-pass filter in rads/s. The significant part of the spurious signal as given in Equation (37) is

$$\omega_q = \frac{\sin n\frac{\pi}{2}}{n\frac{\pi}{2}} \cos\ (\omega_m - n\omega_s)t \qquad (39)$$

In Equation (39), since the ratio of modulating signal to spurious signal is desired, the amplitude of the spurious signal has been normalized to the amplitude of the modulating signal, which is the first term in the braces of Equation (37). When the indicated operations are performed on Equations (38) and (39), the resulting steady-state component is found to be

$$\omega_{oqs} = \frac{\sin\left(n\frac{\pi}{2}\right)}{\left(n\frac{\pi}{2}\right)}\left[\frac{1}{1+\left(\frac{\omega_m - n\omega_s}{K\omega_3}\right)^2}\right]^{1/2} \sin\left[(\omega_m - n\omega_s)t - \psi\right] \quad (40)$$

for K >> 1. From Equation (40), the relative attenuation of the spurious signal is seen to be

$$\gamma = 20 \log_{10}\left(\frac{n\pi}{2}\right)\left[1 + \left(\frac{\omega_m - n\omega_s}{K\omega_3}\right)^2\right]^{1/2} \quad (41)$$

for n odd only. Equation (41) is plotted in Figure 16, which shows that beat frequencies produced between the modulation frequency and harmonics of the switching frequency must be relatively high if significant spurious signals are to be avoided in the output. For example, with $f_3 = 0.01$ Hz and $K = 500$, beats between a modulation frequency and the third harmonic of the switching frequency must be as high as 106 Hz for a 40-dB spurious level and 1060 Hz for a 60-dB spurious level. The effect of beat frequencies produced by the higher-order harmonics of the switching frequency is, of course, less.

The need for a low switching frequency is evident. It is also likely, depending upon the baseband characteristics, that a bandpass filter should be used between the discriminator output and the synchronous detector to limit the response to the fundamental of the switching frequency and the narrow spectum around the fundamental.

Despite this problem the gated AFC system offers several advantages. Among these are freedom from discriminator center frequency stability effects, the need for only one frequency multiplier, and single crystal control—some of the systems to be discussed subsequently in this section require two crystals.

The RF switch used at the input to the limiter discriminator normally would be considered a problem. Recent advances in semiconductor technology, however, have resulted in the availability of PIN surface-oriented diodes designed for use with microstrip in integrated circuitry. These diodes appear to be ideally suited for this switch. In addition, a microstrip approach to the S-band discriminator is available, the bandwidth of which can be made broad enough to cover a telemetry band. Discriminators operating in the low MHz region are not compatible with the fabrication requirements of integrated circuits.

For these reasons, the gated discriminator AFC system will be further investigated in the next phase of this program. Further analysis of the system

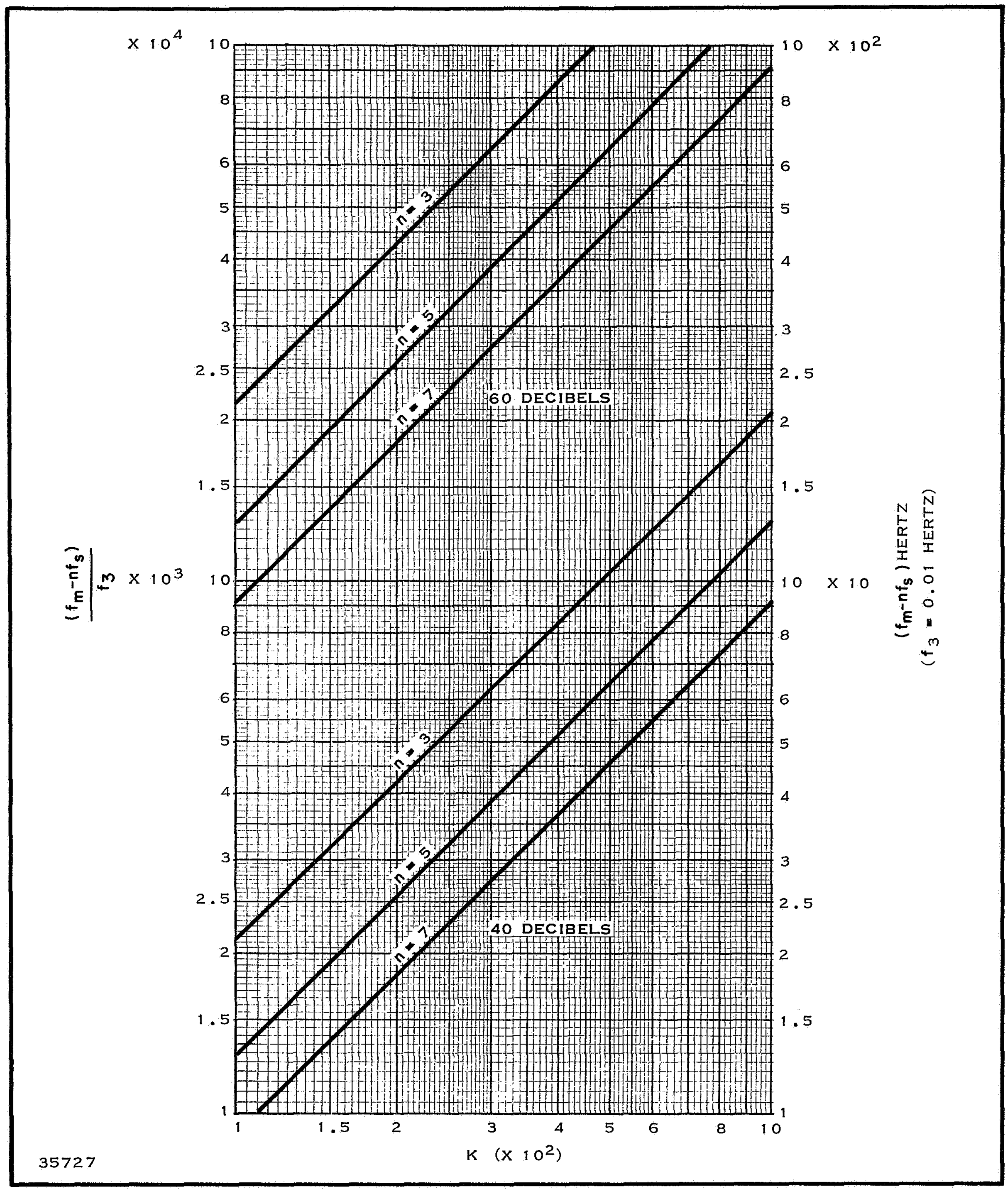

Figure 16. Beat Frequency Requirements To Maintain Spurious Level in the Output 40 dB and 60 dB Below the Level of the Modulation Produced by a Modulating Sinusoid of Peak Frequency Deviation

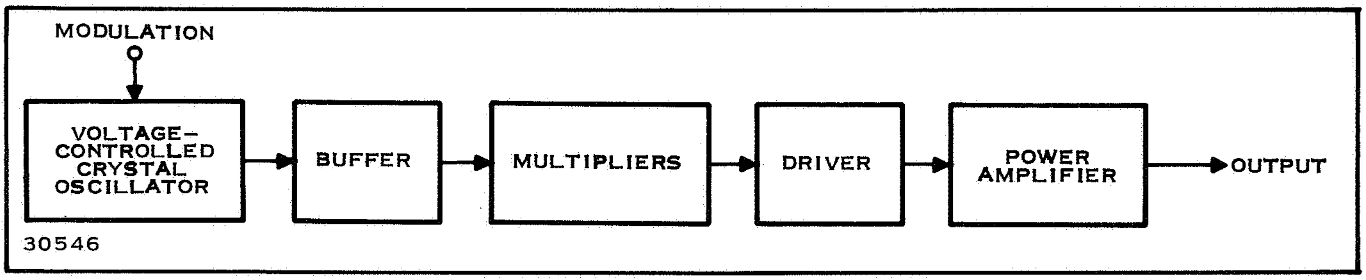

Figure 17. Voltage-controlled Crystal Oscillator FM Transmitter

will be done based on VCO stability studies, which determine the open-loop gain required, and novel techniques of contructing low-pass filters having complex transfer functions designed to provide a stable loop while providing greater attenuation of the low-frequency response. These two factors determine the level of spurious signal generation in the output of the transmitter.

C. VOLTAGE-CONTROLLED CRYSTAL OSCILLATOR

Except for the voltage-controlled crystal oscillator FM transmitter (Figure 17), all of the direct methods of producing frequency modulation require some form of automatic frequency control for achieving 0.005-percent stability. Although the basic stability of the crystal oscillator is degraded when operated as a VCXO, because of the detuning necessary to allow modulation, 0.005 percent is readily obtained. This is especially true when the deviation percentage is small.

A simple method for deviating a crystal-controlled oscillator introduces into the feedback path a series-resonant circuit consisting of a coil and a variable-capacitance diode. The effect of this series circuit is to displace the series resonance above or below the series-resonant frequency of the crystal alone. This phemonema is shown in Figure 18, for which the inductive reactance is larger. The frequency of oscillation is lowered and, thus, further removed from the crystal antiresonant frequency.

The addition of this circuit has created a second reactance zero at f_2. The loop gain at this frequency can be reduced by placing a tuned amplifier in the forward path.

In telemetry applications FM distortion is a very important consideration. At low modulation frequencies the static linearity is a good measure of FM distortion. The frequency response cutoff is determined by the RC network used to apply the modulating voltage to the variable capacitance diode. When the modulating frequency is above a few kilohertz, there is another source of FM distortion due to the spurious response of the crystal. In particular, the crystal must be completely free of unwanted modes in a region corresponding to the generated FM spectrum. As an example, consider a 30-MHz VCXO with a linear characteristic over a ±75-KHz deviation range (Figure 19).

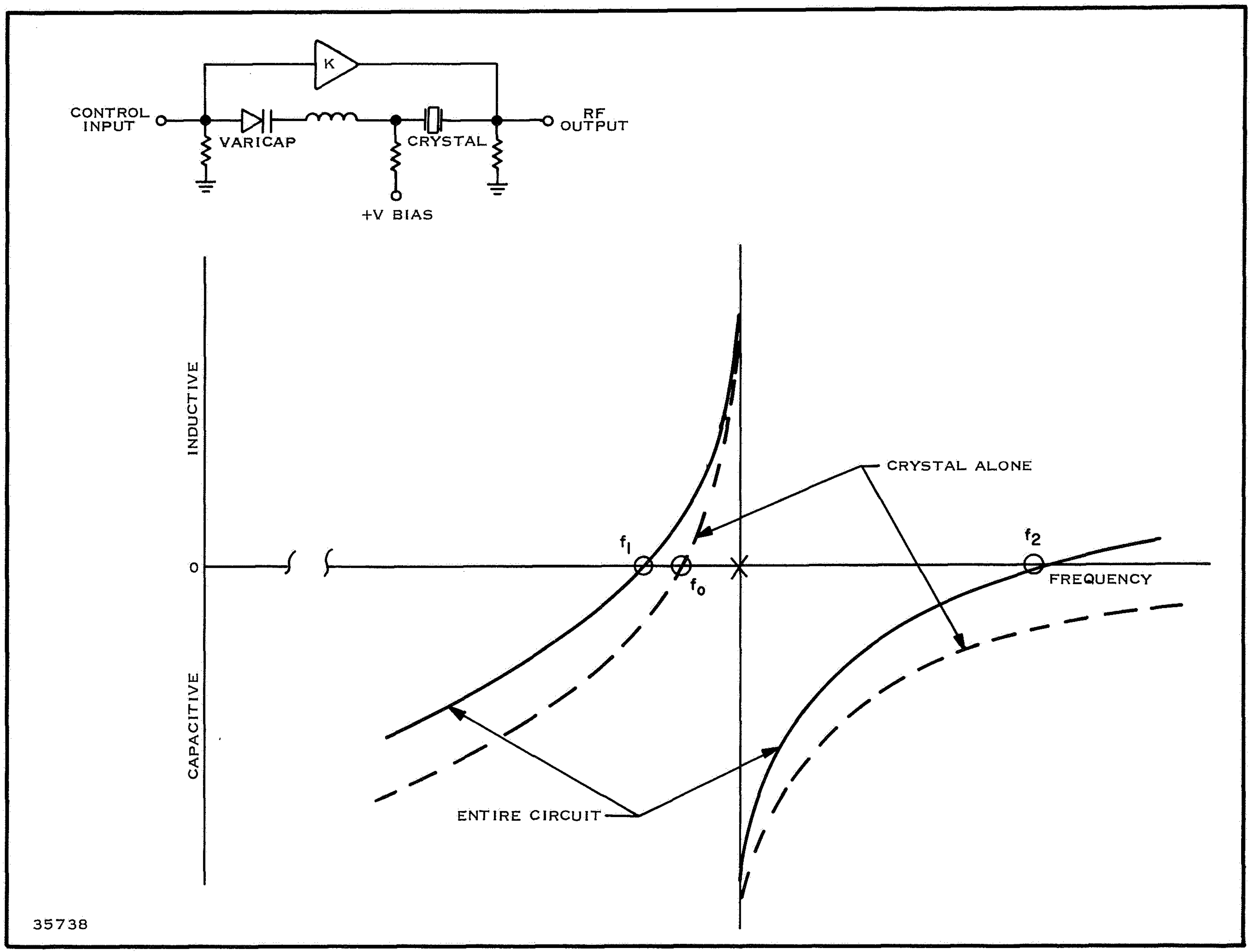

Figure 18. Basic Voltage-controlled Crystal Oscillator Circuit and Reactance Plot

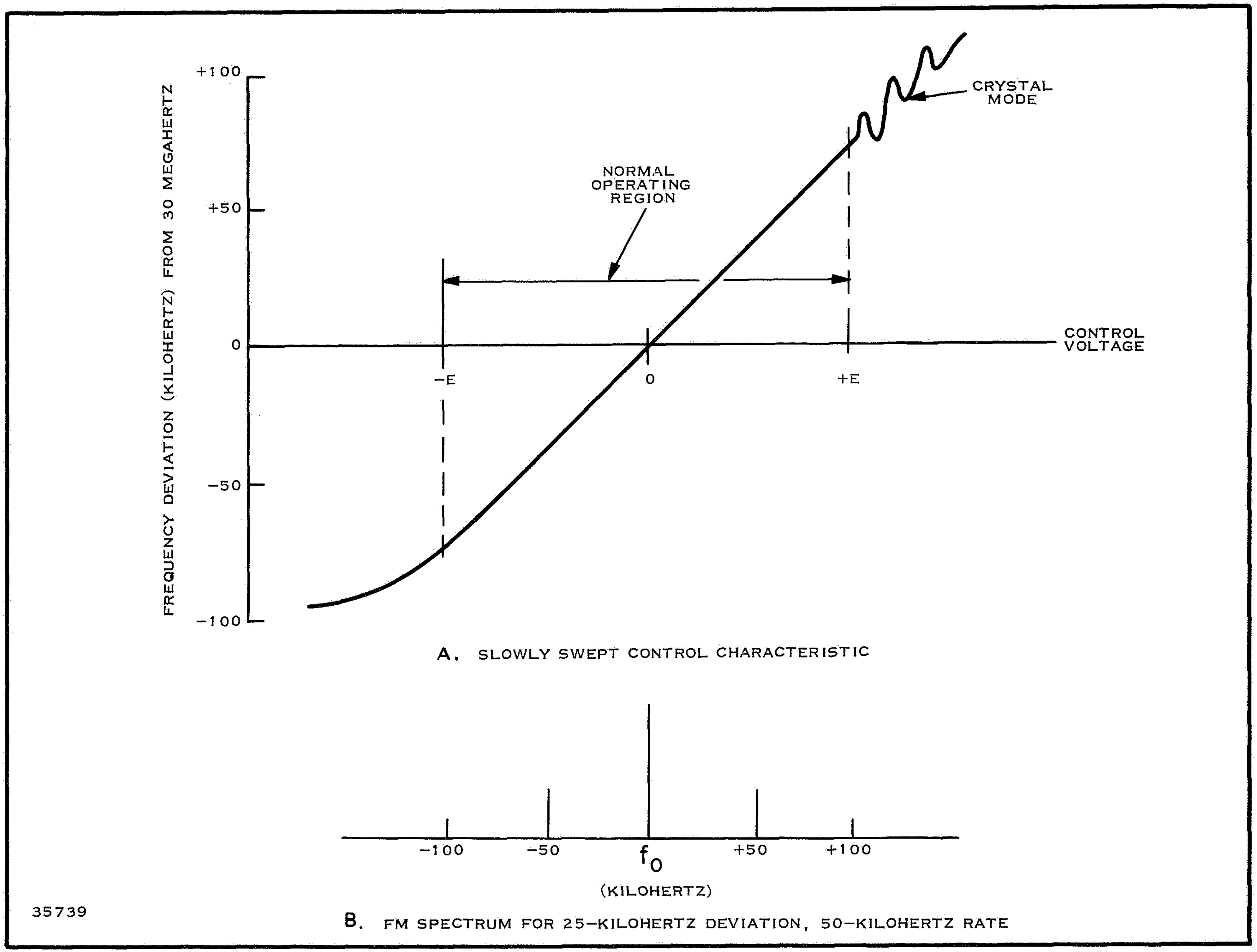

Figure 19. Voltage-controlled Crystal Oscillator Swept Characteristic, Showing Effect of Unwanted Crystal Resonances

If the FM spectrum is confined to this region, the FM distortion will be comparable to the static departure from a straight line. If this unit were to be modulated at a rate of 50 KHz, however, FM distortion will occur as the 100 KHz second-order FM sidebands become significant, for example, for deviations greater than about 25 KHz. FM distortion at high rates depends upon the level of the nearby modes of the crystal and reaches a maximum whenever an FM sideband frequency happens to coincide with the crystal spurious mode.

Operation of the crystal in the region below its series-resonant frequency is the technique used with most VCXO's. As the deviation requirement is increased, it becomes necessary to work the crystal further away from its series-resonant frequency, the result being that the stability suffers. It is important that the frequency not be deviated too close to the series-resonant frequency of the crystal, since in this region the linearity is poor. Less difficulty is encountered with spurious modes when fundamental mode crystals are used rather than overtone crystals. However, the penalty for using fundamental mode crystals is that greater frequency multiplication is required to achieve the final output frequency.

Greater stability is achieved in another way by operating the crystal at its series-resonant frequency. Centering the frequency swing at the series-resonant point results in frequency stability almost equal to the crystal itself, even for relatively wide deviations. With proper crystal and circuit design, superior linearity in addition to high stability can be obtained. The crystal must be designed for VCXO service; otherwise, the modulation will be severely distorted.

Because of spurious modes of operation in the crystal, the RF spectrum must be limited. In the past, the spectrum had to be limited to +50 KHz from the rest frequency of the oscillator. Recent advances by manufacturers of crystals have yielded experimental crystals that can be operated in oscillator circuits at frequencies in the range of 30 to 35 MHz with spectrums as wide as 600 KHz and distortion of no more than 5 percent. These are experimental crystals; production crystals capable of ±300 KHz are readily available.

Even with the experimental units and very small deviation ratios, however, the VCXO approach cannot be used when the baseband response extends to 1 MHz, the current requirement. Regardless of how low the deviation ratio is made, the first-order sidebands will exist out to ±1 MHz, far beyond the deviation rates available.

D. HETERODYNE FREQUENCY STABILIZATION

The heterodyne system (Figure 20) is an old technique that has been used extensively. The stability of the signal at the output of the crystal-controlled oscillator is much greater than the stability of the signal at the output of the

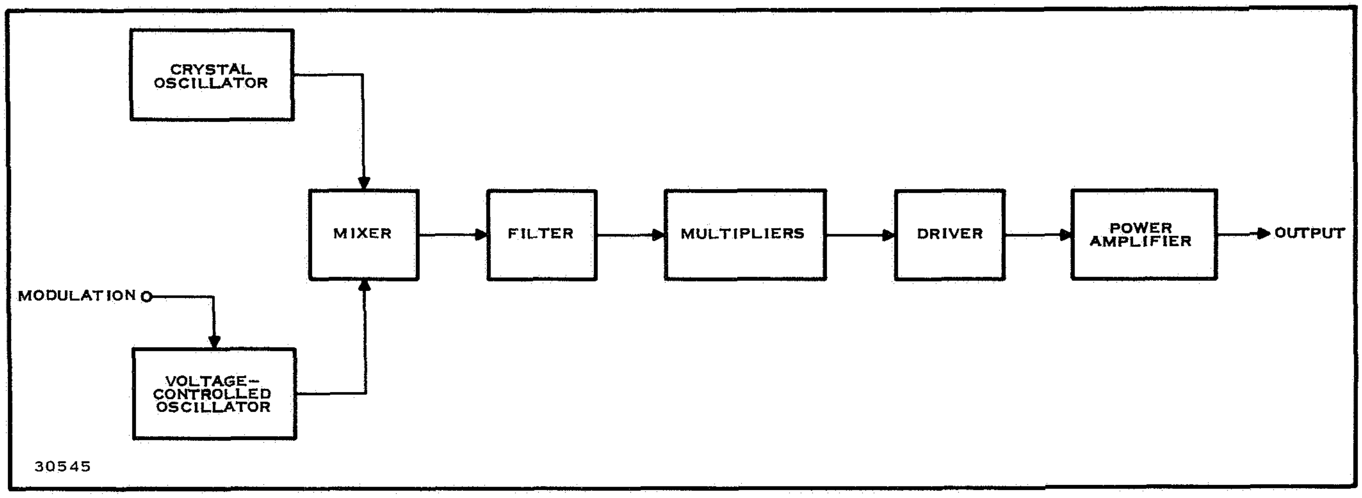

Figure 20. Heterodyne FM Transmitter

VCO. The ratio of these two must be made large so that the output may be made more dependent on the stable source rather than the unstable source.

The output frequency is the multiplied sum of the two sources:

$$\omega_o = M(\omega_c + \omega_v) \tag{42}$$

where

ω_o is the output frequency in rads/s
ω_c is the crystal-controlled oscillator frequency in rads/s
ω_v is the VCO center frequency in rads/s

With Equation (42) the effect of the various center frequency stabilities can be determined with

$$\delta_o \omega_o' = M(\delta_c \omega_c' + \delta_v \omega_v') \tag{43}$$

The primed variables denote design center values; for example,

$$\omega_o' = M(\omega_c' + \omega_v') \tag{44}$$

where

δ_o is the output frequency stability in percent
δ_c is the crystal oscillator frequency stability in percent
δ_v is the VCO frequency stability in percent.

Substituting Equation (44) in Equation (43), we obtain

$$\frac{\omega_c'}{\omega_v'} = \frac{\delta_v - \delta_o}{\delta_o - \delta_c} \tag{45}$$

Equation (45) is plotted in Figure 21 for a typical VCO stability range and for crystal-oscillator stabilities of 0.001 and 0.002 percent. The ratio required is high. Even for a crystal-oscillator stability of 0.001 percent and a VCO stability of 0.5 percent, f_c must be 125 times f_v.

Unfortunately, other considerations also govern the choice of mixing ratio. Since the baseband response extends to 1 MHz, the VCO frequency spectrum will be at least 2 MHz wide and the center frequency of the VCO must be 10 to 20 MHz. The high mixing ratio places the output of the mixer in the 1- to 2-GHz range, eliminating the need for a multiplier following the mixer. But when this is done, the full deviation (1.5 MHz) must be achieved in the VCO. It was previously shown that the bandwidth required for the fully deviated baseband is 8 MHz. The necessary 1 percent linearity can be obtained once again only by raising the center frequency of the VCO. At this point it is interesting to note that with good VCO stability (0.5 percent) the center frequency of the VCO cannot exceed 20 MHz if the output stability of 0.005 percent is to be maintained at 2 GHz. Thus, in this case, the heterodyne technique cannot be used. This basic approach to stabilization is, in general, not useful when the stability of the output is required to be more than an order of magnitude less than the VCO stability.

E. QUADRATURE AFC

The configuration represented in Figure 22 has been used in receivers as a frequency acquisition circuit. The system functions to control the transmitter output frequency to the reference frequency produced by the crystal oscillator and its multiplier. The output of the multiplier is mixed with the incoming VCO frequency in one mixer; a second mixing takes place with the multiplier output shifted 90 degrees. The outputs of the two mixers are in quadrature and contain a difference frequency term and a sum frequency term. The two low-pass filters remove the sum term. One of these signals is then differentiated to produce a signal proportional in amplitude to the difference between the output frequency and the reference frequency, and that signal is then synchronously detected with the output of the other mixer. The output of the product detector contains a dc term and a component at twice the difference of the reference frequency and the transmitter frequency; the latter is removed by the low-pass filter feeding the VCO. The dc term is proportional in amplitude and sign to the desired difference frequency. In Figure 22, the signals existing at those points designated by circled numerals are:

1 $\quad A_o \cos \omega_o t \qquad (46)$

2 $\quad A_r \cos (\omega_r t + \theta) \qquad (47)$

3 $\quad A_r \sin (\omega_r t + \theta) \qquad (48)$

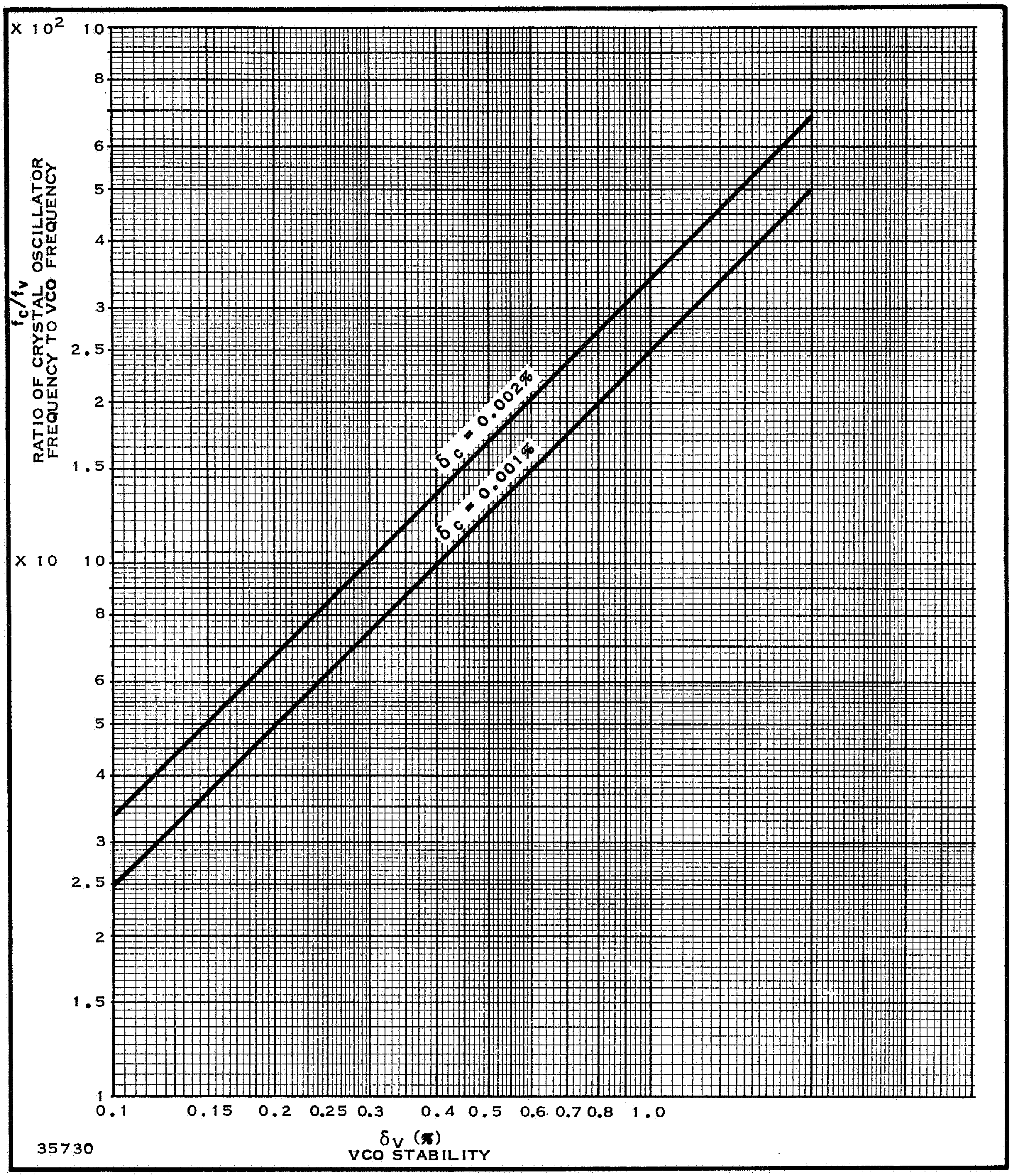

Figure 21. Ratio of the Crystal-controlled Oscillator Frequency to the VCO Frequency as a Function of the VCO Stability for Two Crystal-oscillator Stabilities

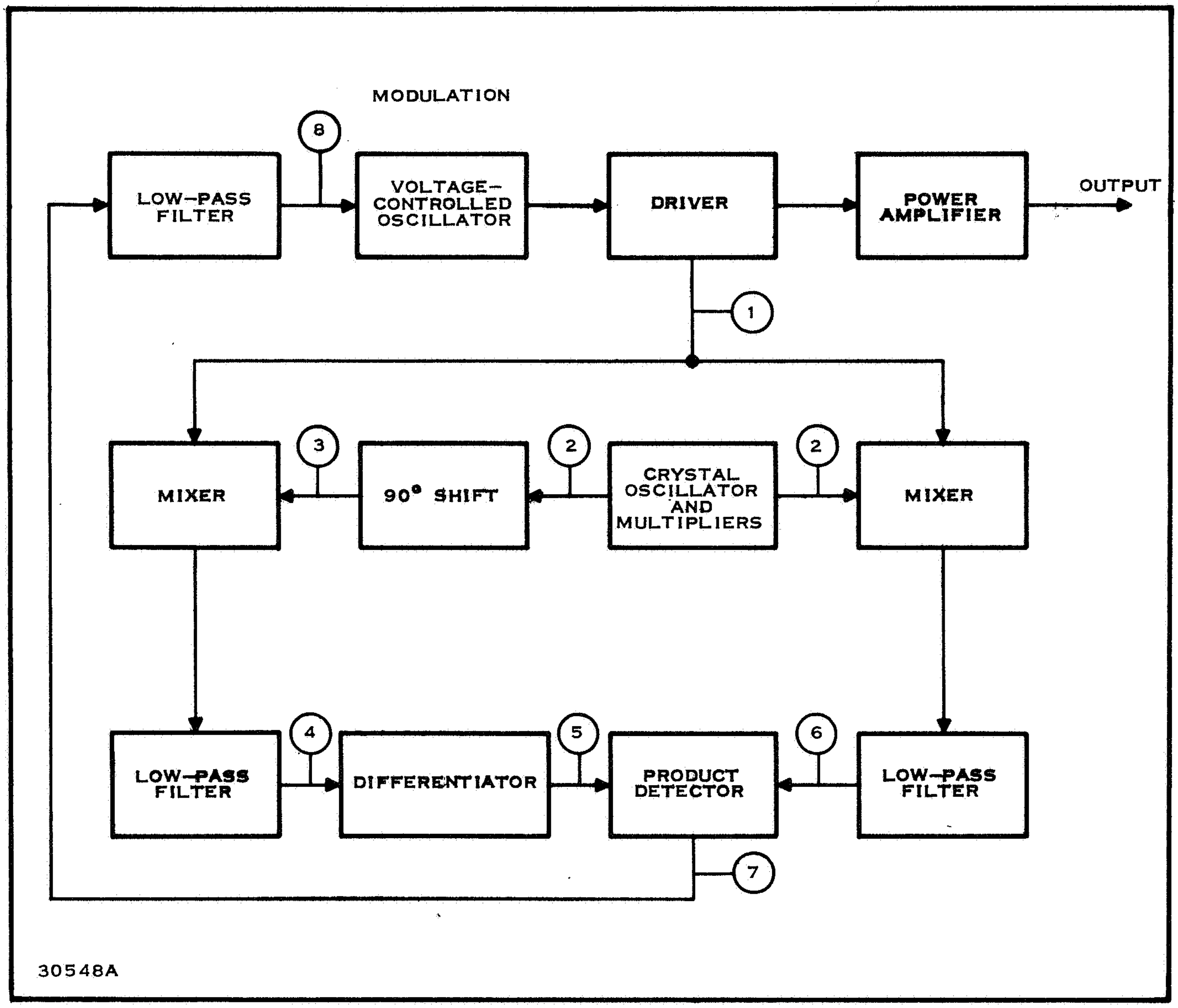

Figure 22. Quadrature AFC—FM Transmitter

④ $$\frac{A_oA_r}{2}\sin[(\omega_r - \omega_o)t + \theta] \tag{49}$$

⑤ $$\frac{A_oA_r}{2}(\omega_r - \omega_o)\cos[(\omega_r - \omega_o)t + \theta] \tag{50}$$

⑥ $$\frac{A_oA_r}{2}\cos[(\omega_r - \omega_o)t + \theta] \tag{51}$$

⑦ $$\left(\frac{A_oA_r}{2}\right)^2(\omega_r - \omega_o)\left\{\frac{1}{2} + \frac{1}{2}\cos 2[(\omega_r - \omega_o)t + \theta]\right\} \tag{52}$$

⑧ $$\left(\frac{A_oA_r}{2}\right)^2\frac{(\omega_r - \omega_o)}{2} \tag{53}$$

The major problems in this system center around implementation of the differentiator and maintenance of the 90-degree phase shift through the system to the product detector. For telemetry systems the total allocation in a given frequency range is on the order of 5 to 10 percent. A branch-line hybrid (easily constructed in microstrip form) can be used for bandwidths this narrow, allowing operation anywhere in the band. Phase shift, nonetheless, can be introduced into the system in the mixers and other components. Fortunately, the system degrades slowly for small variations in the phase shift; that is, the output as given by Equation (53) is multiplied by the cosine of the error phase angle, the value of which is close to unity for small error angles. At the required bandwidth, the differentiator must produce the derivative of the input signal over a frequency range of three decades, which is the range of the uncompensated VCO frequency to its compensated value. The three decades of frequency range produce a voltage range of 60 dB at the output of the differentiator. A brief investigation of integrated circuit operational amplifier differentiators has shown that the noise level can be held to about 90 dB below the maximum output, thus providing a sufficient margin over the requirement.

This particular system has the fundamental capability of meeting the basic transmitter specification for frequency stability. Its problems center around the difficulty of implementing the circuit functions required. For this reason, this system will be investigated further in the next phase of the program.

F. PULSE DISCRIMINATOR AFC

The pulse discriminator AFC system (Figure 23) is basically a pulse circuit implementation of the quadrature AFC system just discussed. As shown in Figure 23, some of the transmitter output is fed to the buffer amplifier, which in turn applies this signal to the two mixers. The crystal oscillator drives the multipler, the output of which is the reference frequency. This reference signal is also applied to the two mixers, one of the signals first being passed through a 90-degree phase-shift network. In the output of the mixers, the sum terms are filtered, leaving only the difference frequency terms. Limiting operations are then performed on these two outputs.

The phase splitter and the differentiator develop pulses at the positive and negative zero crossings of one of the limited signals. These two pulse trains are summed separately with the other limited signal to produce the outputs shown as E and F in the diagram. As a result of the bias level set on the diode summing junction, and the phase relationship of the difference frequency outputs of the mixers, one or the other of the signals at E and F contains pulses which exceed the bias level. Which of the two outputs is active is determined by the sign of the difference frequency, for example, whether the reference frequency is above or below the VCO frequency. The frequency of the output pulse rate is equal to the difference between the reference and VCO frequencies. The pulses are then integrated to provide

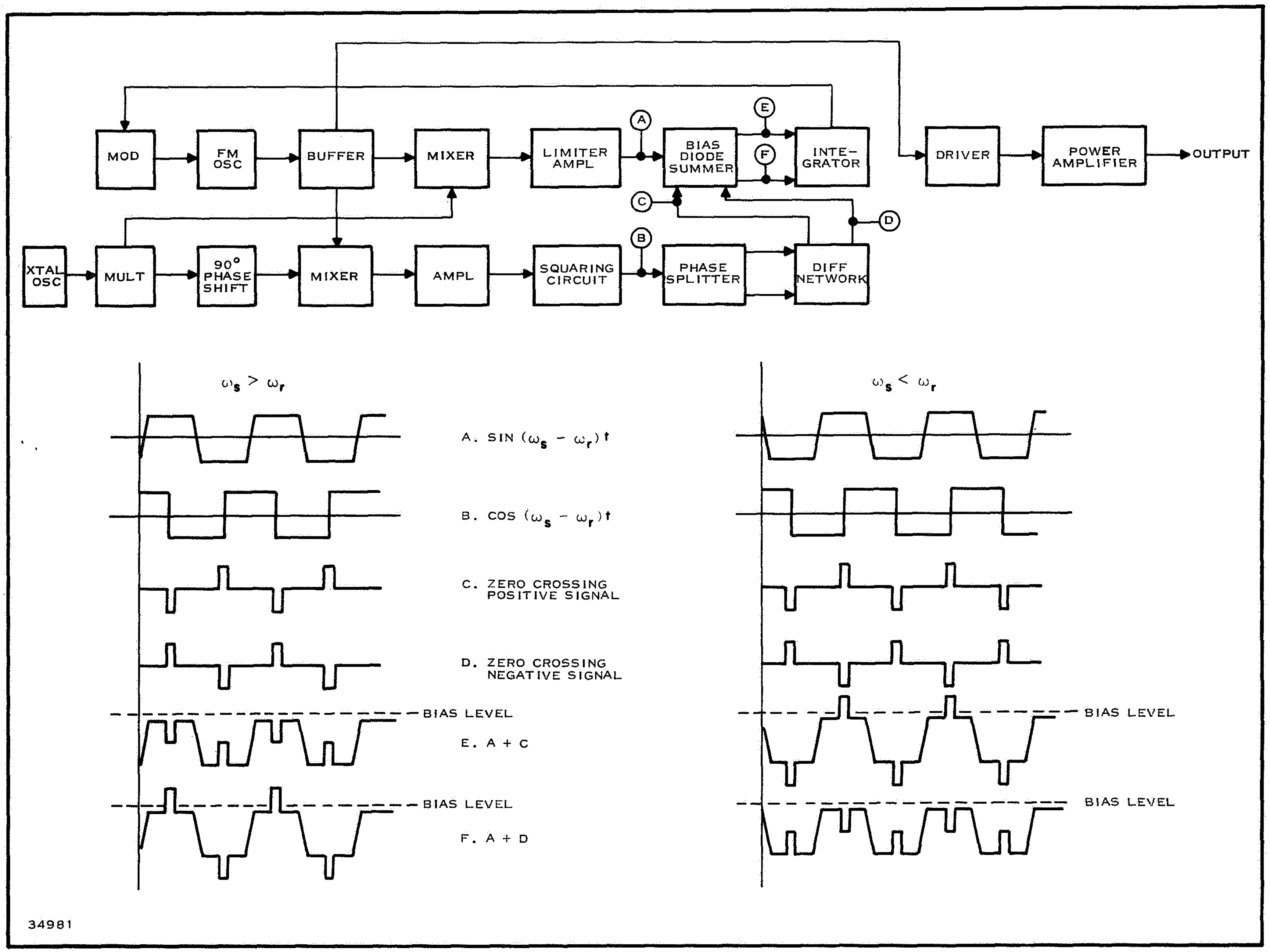

Figure 23. Pulse Discriminator AFC—FM Transmitter

an error signal proportional to the difference frequency, with the polarity determined by the sign of the difference.

Since difference frequencies on the order of 20 to 50 MHz will be present at the output of the mixers, video bandwidths of 500 MHz will be required for circuitry having the square waves and the differentiated zero crossings of the square waves. This is, of course, a transient condition that exists only when the transmitter turns on, but it is obviously an important condition. After the initial transient condition the AFC loop is in operation and the difference frequency is greatly reduced. At this time the difference frequency is relatively low and actually falls within the same band of frequencies as the modulation. Under these conditions, noise produced by the pulses can be picked up by other circuitry and also produce modulation.

When this system is compared to the quadrature AFC system described previously (the two are conceptually similar), the pulse discriminator is seen to require wide video bandwidths and to operate with small rise-time signals that are capable of producing considerable noise in the range of the baseband. Neither problem exists in the quadrature AFC circuit.

G. DUAL-OSCILLATOR AFC

A wide variety of techniques can be used for center frequency control. In the block diagram of Figure 24, two crystal oscillators and their associated multipliers are used to develop control frequencies equally spaced above and below the desired output frequency. When the VCO frequency is mixed separately with these two control frequencies, difference frequencies are produced which, if the VCO frequency is the correct value, will be equal. With the VCO off frequency, the difference frequencies at the outputs of the mixers are unequal; the difference frequencies are then limited to the same voltage level and differentiated to produce a train of pulses. The difference frequencies are represented by the pulse repetition rate. Both positive and negative pulses are present at the output of the differentiator; the negative pulses are removed by the detector and the two separate positive pulse trains are fed to a difference amplifier. The output of the difference amplifier is then integrated or low-pass filtered to produce a control signal proportional to the difference between the frequencies of the two pulse trains. With the output zero, the frequencies of the two pulse trains will be equal and the average VCO frequency exactly centered between the two crystal-controlled frequencies.

The AFC control characteristic for this system is shown in Figure 25. Note that, unlike a discriminator control characteristic, the output does not fall back to zero volts for frequencies beyond the discriminator peaks; instead, a control signal remains, but proportional control is lost. Proportional control is maintained for VCO frequencies off the design center value of plus and minus one half the difference of the two reference frequencies as shown in Figure 25.

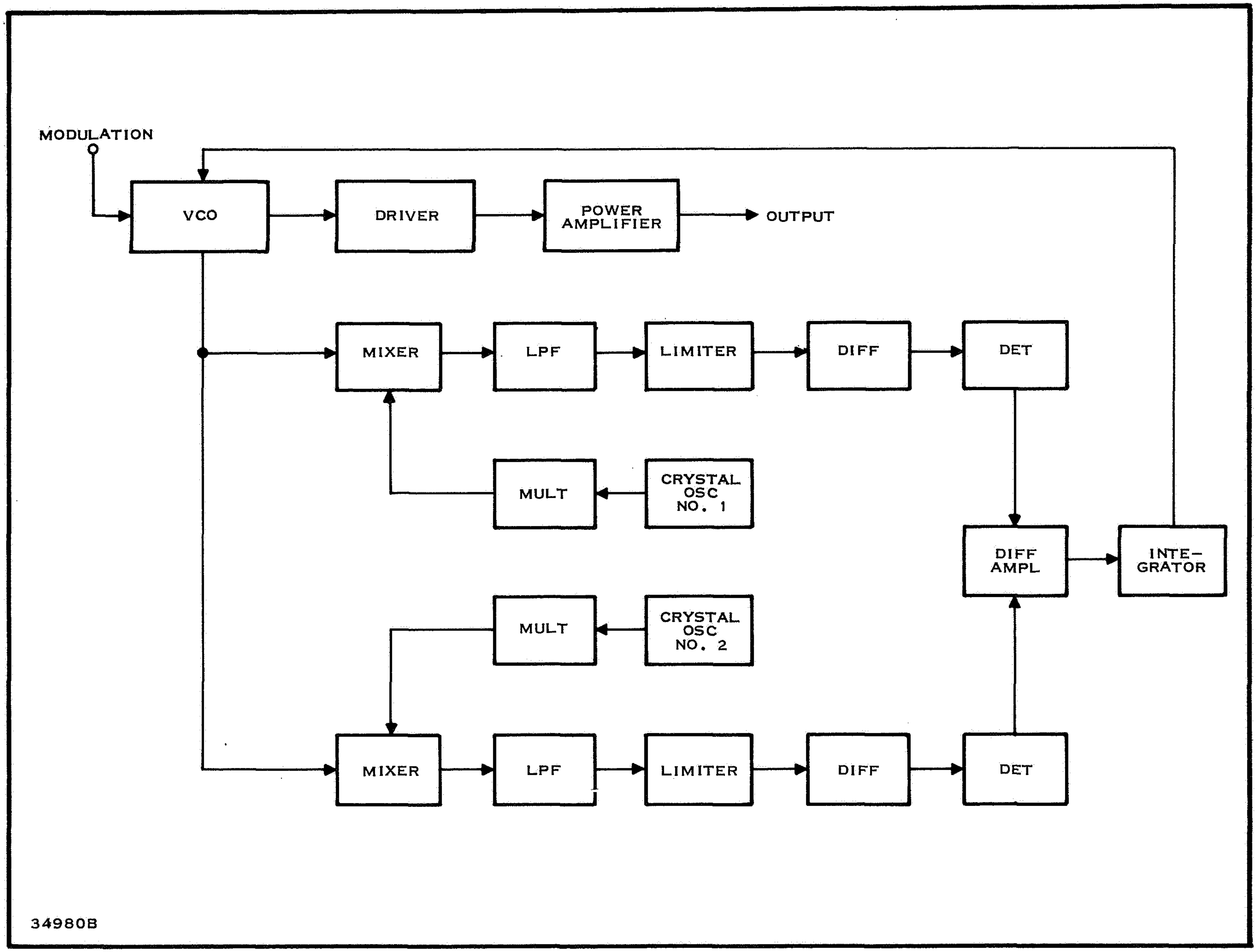

Figure 24. Dual Oscillator AFC—FM Transmitter

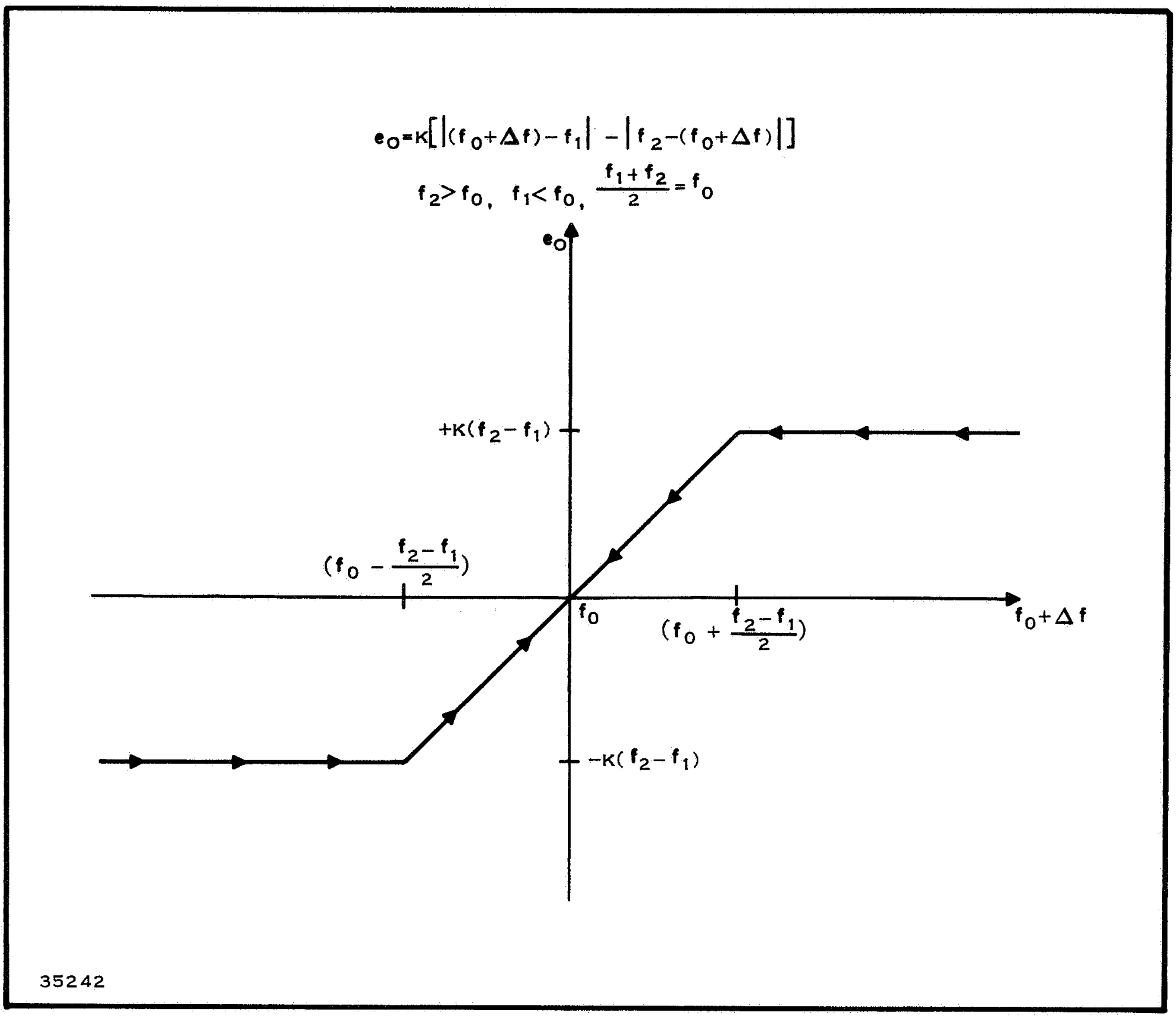

Figure 25. AFC Characteristic for Dual Oscillator AFC System

This system, although conceptually simple, has a number of drawbacks. First, two crystals are required to control the reference frequencies; a single crystal could be used, but the multiplication ratios required would be very large. Next, two multipliers are needed and these are difficult in integrated circuitry so systems using only one multiplier are preferred. Since the multipliers will be operating near the same frequency, the percentage difference is small, and they must be well isolated from each other. Finally, the limiter, differentiator, detector chains must be nearly identical for accurate control of center frequency. Small differences in the limiting amplitudes of the two limiters, the time constants of the differentiators, or the thresholds of the detectors will produce an error in the center frequency control point.

In spite of these problems this system will be investigated from a circuitry standpoint in the next phase of the contract.

H. DUAL-OSCILLATOR SINGLE-MIXER AFC

Two of the problems associated with the previously discussed dual-oscillator AFC are eliminated in the configuration of Figure 26. The functioning of this system is similar to that of the dual-oscillator AFC. In this case, only one difference frequency is produced; the reference frequency is controlled by crystal oscillator number 1. Crystal oscillator number 2 produces a second reference frequency which is equal to the difference between the correct output frequency and the reference frequency. In this arrangement, a multiplier and mixer have been eliminated and only one of the crystals need be changed with center frequency change. Since only one difference frequency is produced and it is referenced to a fixed "bias" frequency, the system has regions of positive feedback as shown in Figure 27. These occur on either the high side of the output center frequency or the low side, depending upon whether the reference is above or below the center frequency, respectively. Proportional control is available for VCO frequencies which are off center frequency by no more than the difference between the reference frequency and the output center frequency. This frequency difference is also equal to the frequency of crystal oscillator number 2.

Although this system has eliminated some of the objections to the dual oscillator AFC, the requirement for balanced limiters, differentiators, and detectors remains. In addition, an undesirable AFC characteristic has been introduced. Other configurations offer greater promise and do not require two crystals, which are large components compared to the other components in microwave integrated circuits.

I. DUAL-OSCILLATOR GATED AFC TYPE I

The system of Figure 28 is essentially a gated version of the dual-oscillator AFC system of Figure 24. The square-wave generator is operated at a very low frequency and performs the function of sampling the two frequency differences that alternately occur at the output of the mixer. A second switch, also controlled by the square-wave generator, gates the pulses at the output of the detector alternately into the two inputs of the difference amplifier. In this way the output of the integrator following the difference amplifier is a control voltage proportional to the difference between the VCO frequency and the average frequency of the two reference frequencies.

The improvement in this system over the dual-oscillator system of Figure 24 is in the elimination of a mixer, limiter, differentiator, and detector. This is a significant improvement, since the balancing of these two chains in the dual-oscillator AFC is a major disadvantage of that system.

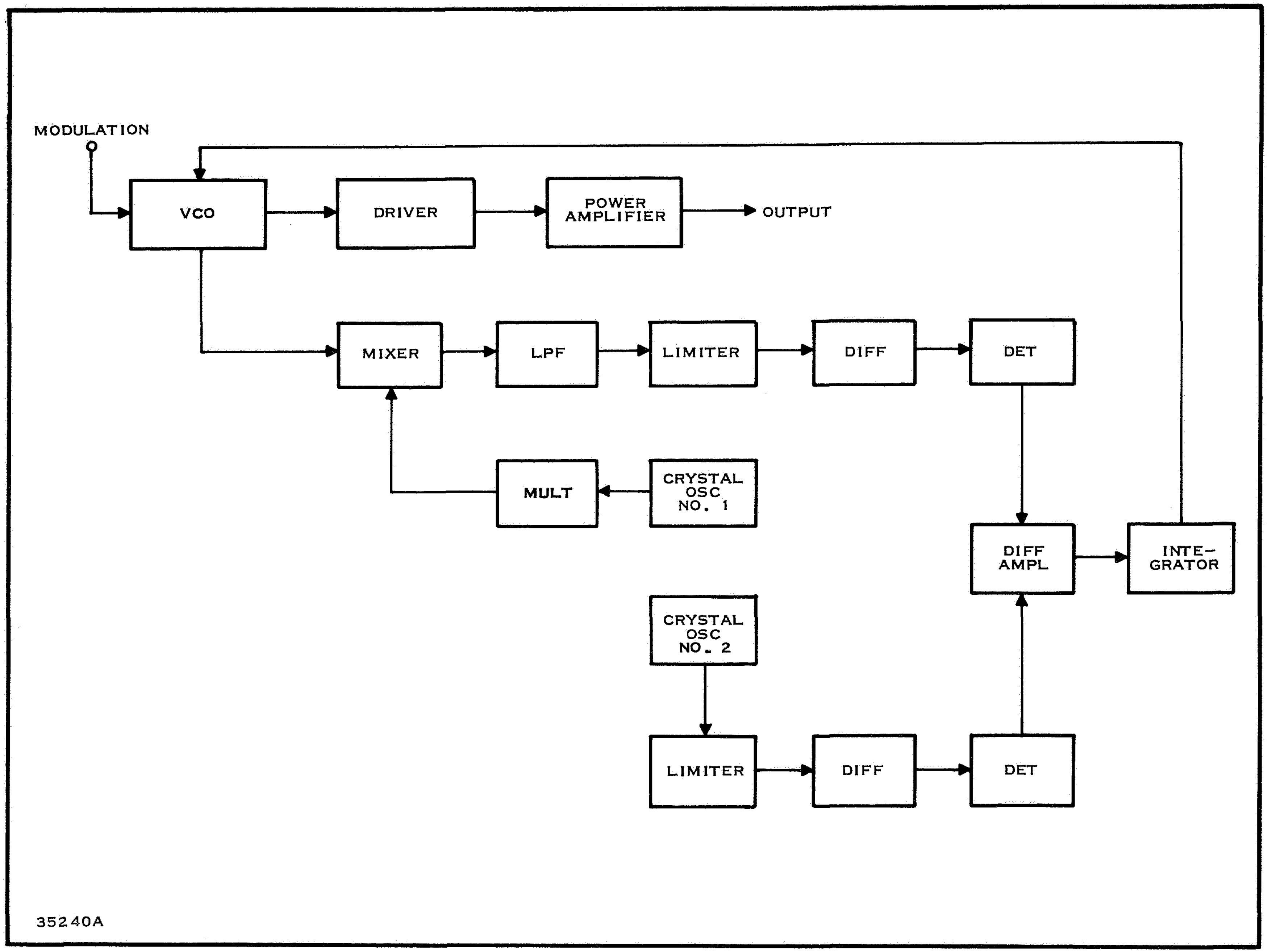

Figure 26. Dual Oscillator Single Mixer AFC—FM Transmitter

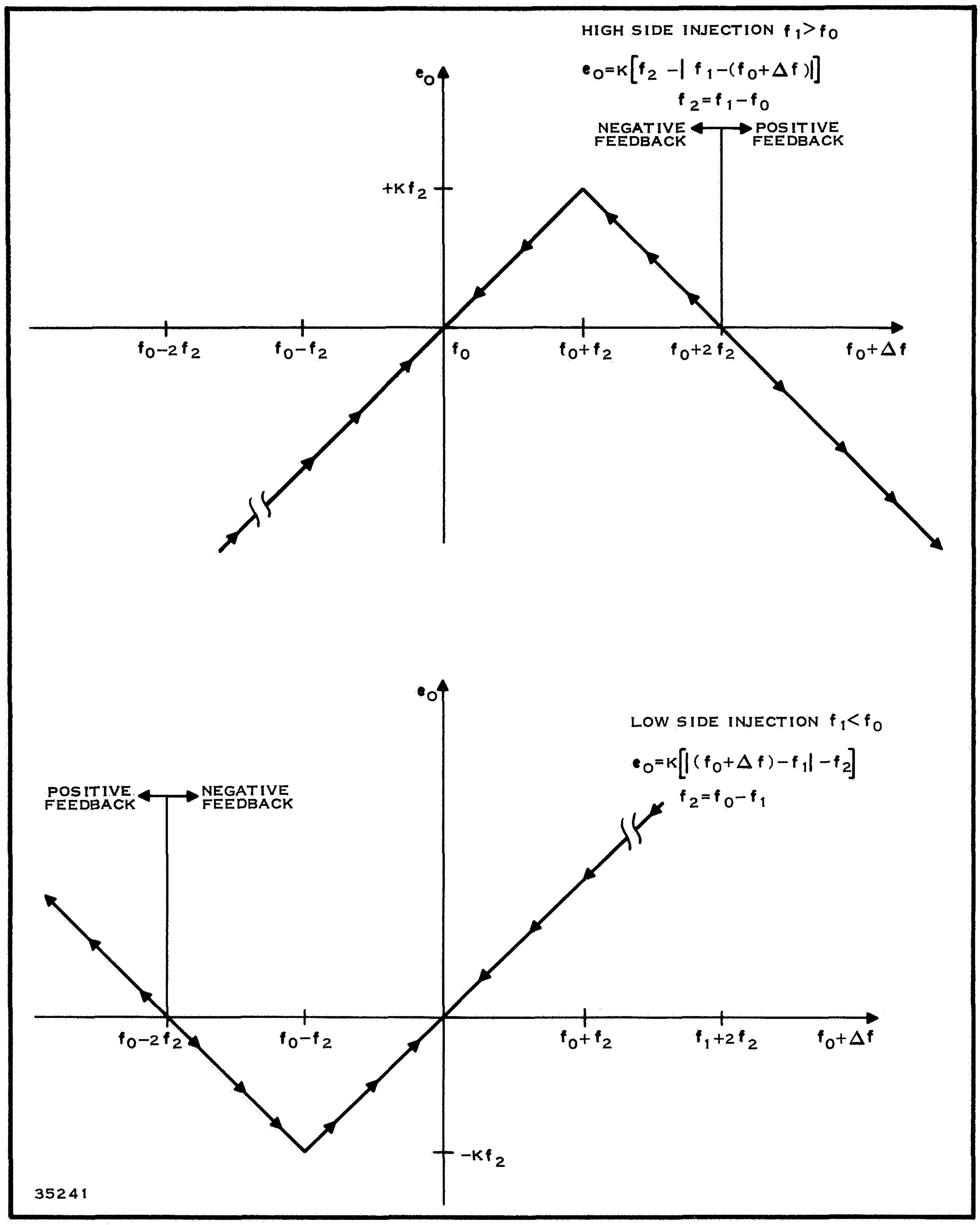

Figure 27. AFC Characteristic for Dual Oscillator Single Mixer AFC System

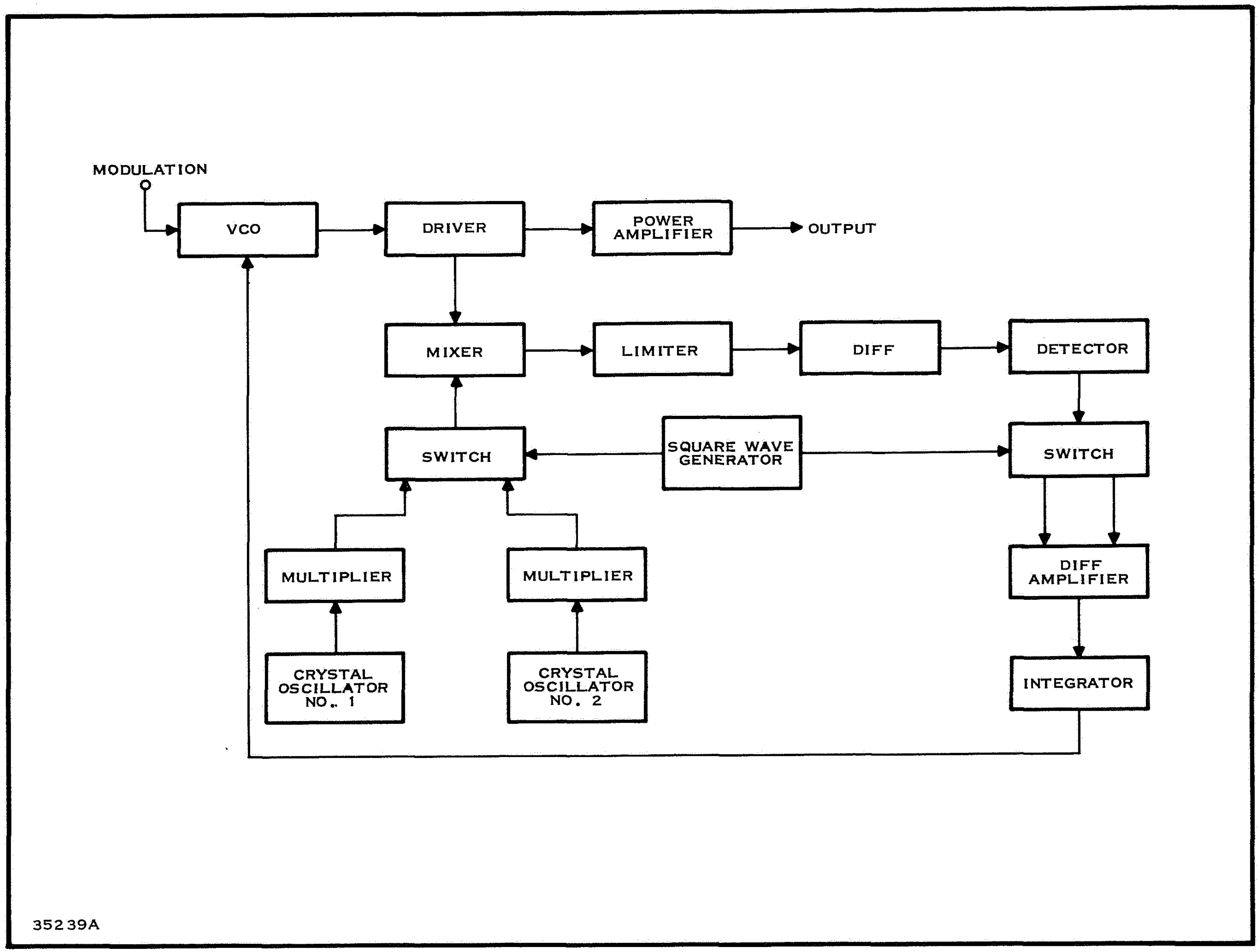

Figure 28. Dual Oscillator Gated AFC, Type I—FM Transmitter

Still, the requirement for two crystal-controlled oscillators, both of which must be changed with changes in the assigned frequency, and two multipliers is a disadvantage.

In addition, spurious signal generation exists due to harmonics of the switching frequency beating with the lower modulation frequencies. This problem is discussed in Section IV. B in connection with the gated discriminator AFC system. This system may be compared with the gated discriminator AFC system of Figure 12, and when this is done, it is seen that the discriminator used in that system has been eliminated at the cost of a mixer and an additional multiplier and crystal oscillator.

J. DUAL-OSCILLATOR GATED AFC, TYPE II

In the system of Figure 29, one of the multipliers has been eliminated. Crystal oscillator number 2 operates at a low frequency equal to the difference of the output center frequency and the reference frequency at the output of the multiplier driven by crystal oscillator number 1. This system is a gated version of the dual-oscillator single-mixer AFC system shown in Figure 26. Its advantage over the system of Figure 26 is that it does not require the balancing of two limiter, differentiator, and detector chains. Its advantage over the Type I system, discussed in the previous subsection and shown in Figure 28, is that one multiplier has been eliminated and only one crystal need be changed with changes in the assigned frequency. The spurious problems associated with the switch remain, however, and the poor AFC control characteristic shown in Figure 27 and discussed in Section IV. H also applies here.

K. VOLTAGE-CONTROLLED CRYSTAL OSCILLATOR AFC

Although a voltage-controlled crystal oscillator cannot be used for the modulator-frequency control function in this transmitter as discussed in Section IV. C (because of the high modulation frequencies involved) it can be used to advantage in an AFC system. This use is shown in Figure 30. In effect the VCXO has replaced the two-crystal system of Figure 28 and the need for two multipliers has been eliminated.

The spurious response problem associated with switching oscillators can be minimized in this system, since it is not necessary to modulate the VCXO with a square wave and thereby generate a spectrum around the two extreme VCXO frequencies consisting of the harmonics of the switching waveform. For example, square-wave integration that can be performed with integrated circuit operational amplifiers will yield a triangular waveform with no even harmonics and with greatly reduced odd harmonics compared to the spectrum of a square wave. This minimizes the amplitude of the beat frequencies occurring between the lower modulation frequencies and the harmonics of the switching waveform. The use of a triangular waveform

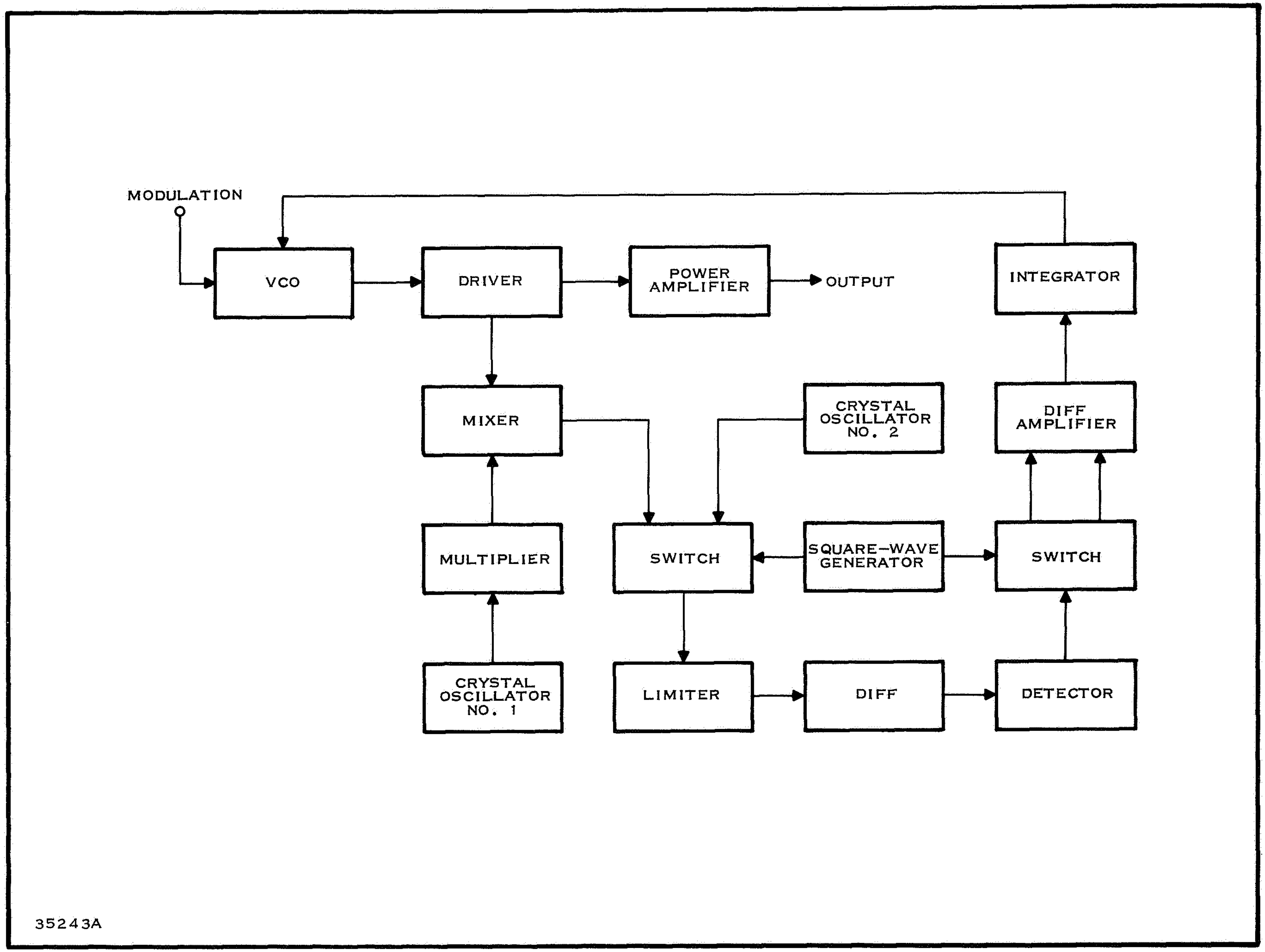

Figure 29. Dual Oscillator Gated AFC, Type II—FM Transmitter

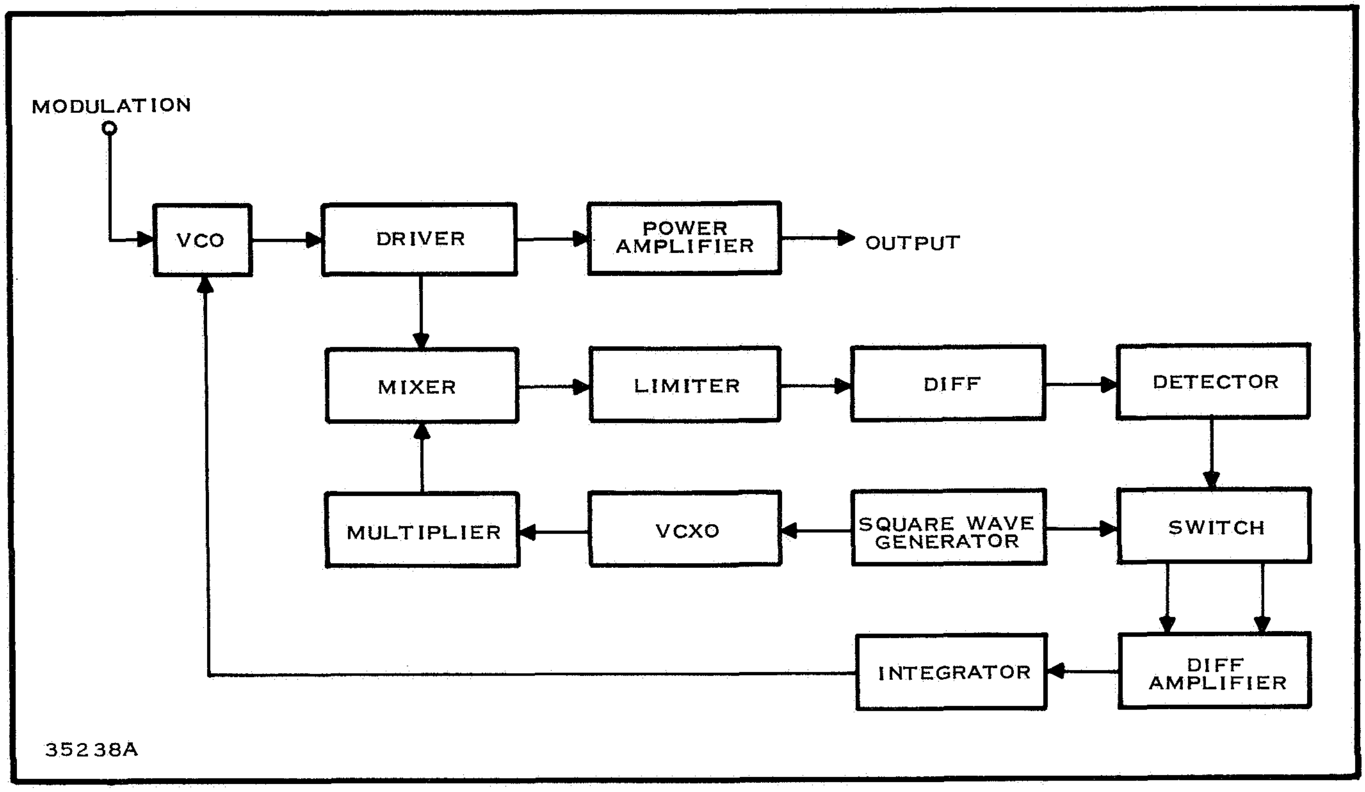

Figure 30. Voltage-controlled Crystal Oscillator Gated AFC—FM Transmitter

for modulating the VCXO also degrades the operation of the AFC loop, primarily in the sense of altering the open loop-gain, which is not a serious objection.

This system has the desirable AFC characteristic shown in Figure 25. To provide an adequate control characteristic when the transmitter is first turned on and the VCO is considerably off frequency, it is necessary to have the peak-to-peak frequency deviation of the VCXO large, on the order of 2 percent. Although great advances have been made in the percentage of deviations attainable with VCXO's in recent years, 1-percent peak-to-peak deviation is about the maximum currently available. Furthermore, the center frequency stability is poor when deviations this large are required. Without oven control, 0.01 percent stability is all that can be provided over any reasonable temperature range. In addition, large percentage deviations are only available with fundamental mode crystals. This limits the crystal oscillator frequency to about 30 MHz and forces greater multiplication than would be required with other AFC systems that can use overtone crystals.

In the next phase of the program, further investigation of this use of VCXO's will be made. Probably, the poor center-frequency stability will remain a problem and the system will not prove useful in this application.

SECTION V
PROGRAM PERSONNEL

Name	Title
Albert E. Mason, Jr.	Project Engineer
Louis I. Farber	Engineer

SECTION VI
CONCLUSION

The next study task to be performed is concerned with the analysis of the technical parameters of the various components and techniques used in the FM telemetry transmitter systems selected under this study task. More than one basic system is being carried forward for analysis in the next phase of the program. It is expected that the selection of one system will be accomplished in a short time by concentrating the investigation on specific trouble-spots in the designs. To avoid pre-emption of the results of the next study task, no attempt was made to eliminate all but one of the basic systems in this phase of the program. The results of this study, however, have minimized the number of circuitry investigations that will be required and have prevented the possibility of a detailed circuitry investigation of a basic system that would not, regardless of the implementation, meet the performance specifications.

SECTION VII

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LIST OF SYMBOLS

A_m	amplitude of detected sinusoid
BW	bandwidth of filter
e_{dc}	dc value of the error voltage
e_ϵ	discriminator output, error voltage
f_b	bit frequency, bits/s
f_h	highest modulating frequency, Hz
f_m	modulating frequency, Hz
G(s)	transfer function, low-pass filter
K_1	VCO constant, radians/second/volt
K_2	discriminator constant, volts/radians/second
K_3	dimensionless: includes gain/loss of amplifiers, frequency converter, filter and summing junction
M	multiplying factor
$M_f(t)$	modulated wave
$M_p(t)$	phase-modulated wave
n	order of harmonics
S(t)	square-wave switching function
V(t)	modulating signal
α	amplitude distortion
β	phase angle
$\beta\ell$	maximum phase shift
γ	relative attenuation
ΔF	peak frequency deviation
$\Delta\omega$	peak deviation
δ_d	discriminator center frequency stability, percent
δ_o	output frequency stability, percent
δ_r	reference oscillator center frequency stability, percent
δ_v	VCO center frequency stability, percent

$\phi(t)$	instantaneous phase difference
Φ	peak phase deviation
ω_3	3-dB cutoff frequency for low-pass filter, equal to 1/RC
ω_c	crystal-controlled oscillator frequency, radians/second
ω_c'	design center frequency for crystal-controlled oscillator, radians/second
ω_h	highest modulating frequency, radians/second
ω_i	instantaneous frequency, radians/second
ω_ℓ	lowest modulating frequency, radians/second
ω_m	modulating frequency, radians/second
ω_o	output frequency, radians/second
ω_o'	design center output frequency, radians/second
ω_{ops}	steady-state component of the instantaneous frequency due to modulation
ω_{oqs}	steady-state component of spurious modulation signal
ω_p	open-loop component of the instantaneous frequency due to modulation
ω_q	spurious modulation component
ω_r	reference frequency, radians/second
ω_s	frequency of spurious signal, radians/second
ω_s	square-wave switching frequency, radians/second
ω_v	VCO center frequency, radians/second
ω_v'	design center frequency for VCO, radians/second
Ω_d	transform of center frequency discriminator, radians/second
Ω_o	transform of output frequency, radians/second
Ω_{op}	transform of closed-loop component of instantaneous frequency due to modulation
Ω_{oq}	transform of spurious signal component in output
Ω_p	transform of open-loop component of instantaneous frequency due to modulation

Ω_q	transform of spurious modulation component
Ω_r	transform of reference frequency, radians/second
Ω_v	transform of center frequency VCO, radians/second